CHAPTER 23

RADIO FREQUENCY AMPLIFIERS

By B. Sandel, A.S.T.C.

SECTION 1 : INTRODUCTION

(i) Aerial coupling   (ii) Tuning methods   (iii) R-F amplifiers   (iv) Design considerations.

(i) Aerial coupling

In any radio receiving system it is necessary to have some means of effectively transferring the modulated carrier voltage from the aerial system to the grid of the first radio frequency voltage amplifier. The usual method of achieving this result is to use some form of aerial coupling transformer which will give the desired voltage a larger amplitude than that of any other radio frequency voltages which may also be present at the receiver input terminals.

There are many possible forms of aerial coupling arrangements, but the most common type is the inductively coupled transformer, consisting of a primary and a secondary winding. The tapped inductance, or auto-transformer, is also extensively used, particularly with receivers tuning the broadcast F-M band and in portable battery receivers using single turn loops.

The aerial coupling unit is equally applicable to either tuned radio frequency or superheterodyne receivers, since the fundamental problem of effective voltage transfer is the same in each case.

Fixed tuned and untuned aerial stages are sometimes employed. The fixed tuned arrangement is more often used where the desired frequency range to be covered is only a small percentage of the operating frequency, e.g. in a band-spread range of a multi-range receiver, or at v-h-f. The more usual practice is to tune the secondary of the transformer by means of a variable capacitor, or alternatively to make the inductance variable, since this allows the optimum operating conditions, or a close approach to them, to be obtained over the required tuning ranges. It is not generally of great importance whether the tuning is accomplished by using a variable capacitor or a variable inductor. Because of mechanical difficulties with inductance tuners the variable capacitor has been more widely used up to the present time. The inductance tuner is rapidly gaining favour, however, at frequencies around 100 Mc/s, and higher, because of the difficulties due to common metal shafts in ganged capacitors

and the inductance of the metal parts making up the capacitor. With the ganged capacitor arrangement considerable trouble is experienced due to coupling from one circuit to another, because it is not an easy matter to earth the shaft effectively between sections. The earthing wipers or leads have appreciable reactance and resistance at v-h-f and these factors are no longer negligible in comparison with the tuned circuit impedances. Insulated rotors and stators are of considerable assistance in overcoming difficulties due to common shaft coupling.

(ii) Tuning methods

Apart from ganged capacitors, many types of tuners have been used at v-h-f, including vane, guillotine, permeability, resonant line arrangements, etc., but all appear to have some disadvantages and it cannot be said that any one method is greatly superior to all others. Permeability tuning has also achieved some success on the 540 to 1600 Kc/s A-M, broadcast band, but it does not appear likely, in its present form, to supersede the variable capacitor.

(iii) R-F amplifiers

In conjunction with the aerial stage it often becomes necessary to incorporate a further stage to obtain additional radio frequency amplification and to increase further the discrimination against undesired signals, such as those from other broadcasting stations, and in the case of a superheterodyne receiver to limit the effects due to image and other spurious responses. This additional stage usually takes the form of a pentode voltage amplifier operating under class A_1 conditions and using a fixed or variable tuned circuit to act as a load impedance ; an untuned stage may be used in some cases. The valve is connected between the aerial transformer and the r-f tuned circuit which is in turn coupled to a converter valve or another r-f voltage amplifier.

Whether the r-f stage carries out its various functions efficiently depends on a number of factors. Firstly it is necessary that the input resistance of the valve should not be too low at the operating frequency, as this will adversely affect the performance of the input circuit—the aerial circuit in the case of one r-f stage. In addition, the noise generated by the valve should be low, and in a superheterodyne receiver the noise voltage should be very much less than that developed by the converter valve if there is to be appreciable improvement in the signal-to-noise ratio. Also the r-f stage gain should not be too low, otherwise the converter noise will still have an appreciable effect on the signal-to-noise ratio. This last requirement calls for valves operating at v-h-f to have fairly high values of mutual conductance, since the tuned circuit impedances are low and the stage gain is the product of these two factors. Grid-to-plate capacitance should be small in conventional voltage amplifiers to avoid the necessity for neutralization. Input and output capacitances are also important as they set a limit on the possible tuning range. Further considerations of operating conditions for valves used as r-f voltage amplifiers are given in Chapter 2 Sect. 4, and in Sects. 5, 6, 7 and 8 of this chapter.

Since the fundamental principles of operation of the aerial and r-f circuits are very similar whether we use a t-r-f or superheterodyne receiver, it can be taken that the coupling methods and the stage gain and selectivity calculations will apply equally well in both cases. Each type of receiver presents its own particular problems but it is not intended to enter into a discussion of these matters here. Moreover the arguments relating to the relative merits and de-merits of the two systems are well known and details can be found in the references listed at the end of the chapter.

A question which sometimes arises is the desirability of having more than one r-f stage in a superheterodyne receiver. It should be quite clear that, provided the gain of a single r-f stage is sufficient (say 10 to 15 times) to over-ride the effects of converter noise, no appreciable improvement in signal-to-noise ratio is obtained by adding further r-f stages of the same type. The advantage of additional r-f stages lies in the improved selectivity which is possible against image and other undesired frequencies. It is generally preferable to obtain any additional amplification which may be required in the i-f and a-f stages, depending on the purpose for which the

additional gain is required.   Further, a number of tuned r-f stages using ganged capacitors presents a rather difficult tracking problem, as well as giving rise to the increased possibility of unstable operation.   Normally, only the transformer secondary is tuned and the overall selectivity characteristic is not as good as can be obtained with fixed-tuned i-f transformers.   The gain and selectivity of the tuned r-f amplifier varies with the signal frequency, and a fixed intermediate frequency will allow more constant gain and higher gain per stage, for a given range of signal frequencies.   The higher gain per stage is due to the higher dynamic impedance of the i-f transformer when the intermediate frequency is lower than the signal frequency.   Even for an intermediate frequency higher than the signal frequency, larger stage gains are possible because of the better L/C ratio which can generally be obtained.

Even in cases where signal-to-noise ratio and r-f selectivity are not a problem, it is still sometimes advantageous to include a r-f stage.   The r-f stage is used as a buffer to prevent the radiation of undesired signals from the receiver.   A typical case of radiation occurs with a v.h.f. F-M receiver operated in conjunction with an efficient aerial system.   If no r-f stage is used, radiation can occur at the oscillator frequency due to voltages at this frequency appearing at the input circuit of the frequency changer. This is serious because it opens up the possibility of interference between radio receivers which are operating in close proximity to one another.

It is not proposed, in this chapter, to deal in any great detail with grounded grid and other special types of r-f voltage amplifiers, as these are rather specialized applications which are not normally used in the design of narrow-band receivers covering limited bands of frequencies in the range from say 50 Kc/s to 150 Mc/s.   However, at frequencies in excess of about 50 Mc/s, input stages using a grounded-grid triode are generally capable of giving better signal-to-noise ratios than circuits using the conventional grounded-cathode pentode voltage amplifier, but only at the expense of r-f selectivity.   At the lower frequencies the input resistance of the pentode is much higher than that of the grounded grid triode, and so improved gain in the aerial coupling circuit can be used to offset the deterioration in signal-to-noise ratio due to valve noise.   Clearly a better input circuit is possible using a pentode having a lower equivalent noise than with a pentode having the same input resistance but higher equivalent noise, and valve types should be selected with this in mind.

In cases where a wide range of frequencies has to be covered, and where the pass band of the receiver must be very wide (e.g. in television or multi-channel F-M link receivers), the grounded-grid triode offers considerable improvement in signal-to-noise ratio even at frequencies well below 100 Mc/s.   Other applications occur in circuits where neutralization cannot be obtained by normal methods or where the circuit requirements are such that the grid must be earthed.

A special circuit (Cascode) which is of interest (see Ref. A10 Chapter 13) uses a grounded-cathode triode voltage amplifier, which is neutralized, followed by a grounded-grid triode.   This circuit has high input resistance and low equivalent noise.   It offers appreciable improvement in signal-to-noise ratio over the other two types of circuits, particularly at frequencies in excess of about 70 Mc/s.   The combined circuit can be considered as equivalent to a single grounded-cathode triode having zero plate-grid capacitance, and a mutual conductance and noise resistance equal to that of the first triode valve (the valves used can be types 6J4 or 6J6 etc.).

Triode circuits of the types mentioned should not be used blindly, particularly at the lower frequencies, because the plate resistances of triodes are extremely low when compared with those of pentodes, and so the r-f stage gain may be insufficient, because the damping of the tuned circuits will be heavy, and also the r-f stage gain may be insufficient to make the effects of converter noise negligible.   Detailed information on grounded-grid and triode r-f amplifiers can be found in Refs. B8, B18, A10 and A4.

(iv) Design considerations

The design methods to be discussed are equally applicable over any part of the range of frequencies mentioned above, but their practical application is rather more difficult as the frequency becomes greater, mainly because of the physical size of the com-

ponents (including valves) and because the circuit layout is increasingly important if satisfactory results are to be obtained. Electron transit time effects in valves become of greater importance as the frequency is raised because of the changes which occur in the valve input and output admittances*. The valve input admittance must be considered when designing an aerial or r-f circuit as it will affect the dynamic impedance of the circuit, and consequently the selectivity and gain. Valve sockets and other components will add further damping to the input circuits. The effects of valve noise will also be governed by the dynamic impedances of the input circuits and so the various valve effects must be considered in conjunction with the external circuits. This becomes of greater importance as the frequency increases.

A design difficulty with aerial and r-f stages is in obtaining satisfactory performance over a large tuning range. The three to one frequency coverage normally employed on dual wave receivers for the medium (540—1600 Kc/s) and shortwave (e.g. 6-18 Mc/s) bands presents some problems in regard to tracking, and constancy of gain and selectivity. The L/C ratios obtainable with a variable capacitor giving an incremental capacitance range of at least 9 : 1 are satisfactory on the medium wave range but are rather poor on the shortwave band. For these reasons it is common practice to limit the coverage on shortwave bands, in the better class of receiver, to enable improved all round performance to be obtained ; a frequency ratio of about 1.3 : 1 is usual in the region of 10 to 20 Mc/s, with larger ratios at the lower frequencies. An obvious advantage of multi-band receivers is the greatly improved ease of tuning, apart altogether from other considerations of better performance.

To obtain the reduced frequency coverage with variable ganged capacitors having a capacitance range of about 10-400 $\mu\mu$F, various arrangements of series and parallel fixed capacitors are used. If the circuits are arranged to use a combination of series and parallel capacitors for bandspreading, it is possible to make the tuning follow practically any desired law. Further, if the tuning range is small, the oscillator frequency in a superheterodyne receiver can be above or below the signal frequency and three point tracking can still be obtained using series padders in both the signal and oscillator circuits. These matters will receive further consideration in Chapter 25 Sect. 3 on superheterodyne tracking.

Methods for measuring the value of k, the coefficient of coupling, are given in Chapter 26 Sect. 4(ii)E and (iii)B, C. Methods for measuring the primary resonant frequency of an aerial or r-f transformer are covered in Chapter 37 Sect. 4(ii).

SECTION 2 : AERIAL STAGES

(i) Difficulties involved   (ii) Generalized coupling networks   (iii) Mutual inductance coupling   (iv) Tapped inductance   (v) Capacitance coupling   (vi) General summary.

(i) Difficulties involved

The principal difficulty in the design of an aerial stage is that the type of aerial with which the receiver will eventually be used is, in general, not known. If the aerial impedance were known it would usually be a fairly simple matter to draw an equivalent circuit of the aerial and aerial input stage and to determine the complete performance by solving for the conditions existing in the circuit. When the conditions are not known, however, it is usually best to adopt constants for an aerial system which will have the most adverse effect on the receiver performance and to solve for these conditions. A compromise is then made in the circuit arrangement to minimize the effects of changes of gain, selectivity, and tracking with other circuits in the receiver.

(ii) Generalized coupling networks

Fig. 23.1 shows the generalized form for an aerial coupling unit. E_1 is the voltage induced in the aerial. E_2 is the voltage applied to the grid of the first amplifier stage.

*Valve admittances are covered in Chapter 2 Sect. 8.

Z_1 Z_2 and Z_3 are intended to represent any possible arrangement of components for coupling the aerial to the receiver, where Z_1 includes all impedances external to the

FIG. 23·1 GENERALIZED AERIAL COUPLING UNIT.

receiver input terminals and Z_L is the load impedance. This generalized T section can be made to represent all possible coupling arrangements, since it is well known from circuit theory that any complex network composed of linear bilateral impedances can be transformed, at any one frequency, into an equivalent T network
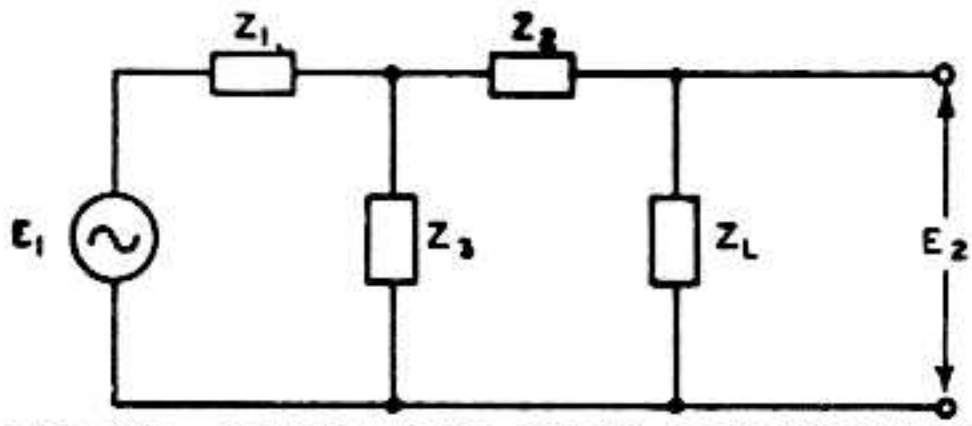
From the generalized network

$$\frac{E_2}{E_1} = \frac{Z_3 Z_L}{Z_1 Z_3 + (Z_1 + Z_3)(Z_2 + Z_L)} \tag{1}$$

This expression is of little assistance unless it is applied to the specialized networks used in radio receivers. Additional generalized expressions for this network are developed in K. R. Sturley's "Radio Receiver Design" Part 1, Chapter 3. It is proposed to deal specifically with a few of the more common types of coupling circuits.

(iii) Mutual inductance coupling

Fig. 23.2A shows one of the most widely used aerial coupling circuits. E_1 is 'the voltage induced in the aerial and Z_A represents the aerial constants, which may be any combination of inductance, capacitance and resistance. L_1 is the aerial coil primary winding which has a r-f resistance R_1. L_2 is the secondary winding coupled by mutual inductance to L_1. R_2 is the r-f resistance of L_2. C_2 is used to tune the secondary circuit to resonance ; it should be noted that the setting for C_2 when the aerial circuit is connected across the primary is different from the setting obtained when L_1 is on open circuit. Fig. 23.2B shows the T section equivalent of the aerial transformer, and Fig. 23.2C shows an alternative equivalent secondary

FIG:23.2(A)

FIG. 23.2(B)

FIG. 23-2(C)　　　　AERIAL TRANSFORMER.

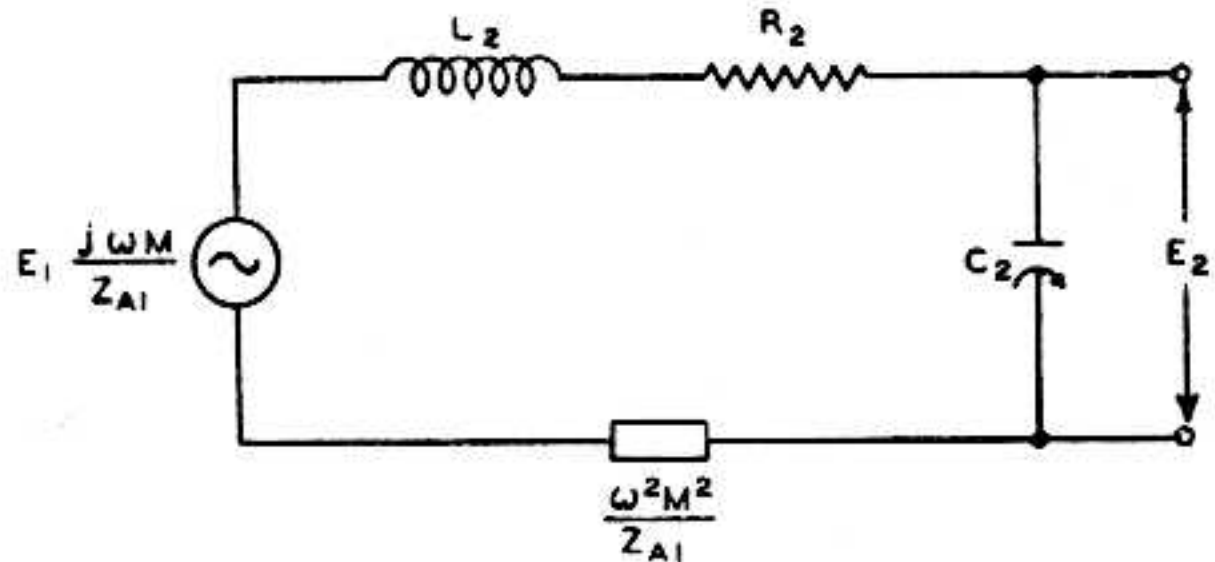circuit which is very helpful in understanding the idea of "reflected" impedances. For the present, the effects of valve loading will not be considered. When the secondary is tuned to resonance, the aerial circuit gain is given by

$$\frac{E_2}{E_1} = \frac{M/C_2}{Z_{A1}\left[R_2 + \dfrac{\omega^2 M^2 R_{A1}}{|Z_{A1}|^2}\right]} \tag{2}$$

where $Z_A = R_A + jX_A$
$Z_1 = R_1 + j\omega L_1$
$M = k\sqrt{L_1 L_2}$

$$Z_{A1} = Z_A + Z_1 = R_{A1} + jX_{A1}$$

and $|Z_{A1}| = \sqrt{R_{A1}^2 + X_{A1}^2}$.

It should be noted that the condition for secondary circuit resonance (partial resonance " S ") is that

$$X_2 = \frac{\omega^2 M^2 X_{A1}}{|Z_{A1}|^2} \tag{3}$$

and the " reflected " resistance and reactance in the secondary are given respectively by

$$\frac{\omega^2 M^2 R_{A1}}{|Z_{A1}|^2} \tag{4}$$

and

$$\frac{-\omega^2 M^2 X_{A1}}{|Z_{A1}|^2} \tag{5}$$

If the mutual inductance M is also made variable, the maximum value of secondary current (and of E_2/E_1 when the primary circuit Q is low) is obtained when

$$M = \frac{|Z_{A1}|}{\omega} \sqrt{\frac{R_2}{R_{A1}}} \tag{6}$$

and

$$\frac{E_2}{E_1} = \frac{1}{2\omega C_2 \sqrt{R_2 R_{A1}}} \tag{7a}$$

$$= \tfrac{1}{2} \sqrt{\frac{R_D}{R_{A1}}}. \tag{7b}$$

where R_D is the dynamic impedance of the secondary circuit. The approximations involved in these equations are discussed in Ref. B12.

From this it can be shown, with M and C_2 adjusted for maximum secondary current, that the selectivity of the tuned circuit is half the value it would have if considered in the absence of the primary. This condition of optimum coupling is not usual in broadcast receivers, except where matching to a transmission line, since the coupling between primary and secondary is made loose to prevent serious tracking errors due to the variable reactance " reflected " into the secondary circuit over a band of frequencies. The expression most generally required is that of eqn. (2) and the calculation of voltage gain need only be made at, say, two or three points across the tuning range.

It should be noted, however, that in some cases loose coupling can lead to a considerable reduction in signal-to-noise ratio if site noise is not predominant, e.g. at the higher frequencies on the short-wave band it is advantageous to use optimum coupling, or slightly greater, to improve the signal-to-noise ratio (see Ref. A2 Chapter 26).

The effects of valve input admittance can be included by considering an additional resistance and a capacitance shunted across C_2. The value of the resistance component may be negative or positive depending on the type of valve and the operating conditions, e.g. some types of converter valves have a negative input resistance.

Having obtained the stage gain it is next important to find the **selectivity** of the circuit. As an approximation for ordinary types of aerial coils it is usually convenient to determine the selectivity of the secondary as for a single tuned circuit. The simplest approach is to consider the resultant Q of the secondary circuit at the resonant frequency when all the secondary resistance, including that due to reflection from the primary and valve loading, is included. The value of Q is given by $\omega L_{eff}/R_{eff}$, where L_{eff} is the resultant secondary inductance. The selectivity is then found as for a single tuned circuit, resonant at the frequency under consideration, using the universal curve given in Chapter 9 (Fig. 9.17) in conjunction with the methods detailed in Chapter 26 Sect. 4(iv)A. This method is satisfactory provided that the amount of mistuning is not too large a proportion of the central reference frequency, and that the aerial coil follows the usual practice with the primary resonated above or below the tuning range or is designed for connection to a transmission line (or aerial) of which the characteristic impedance is largely resistive ; otherwise the variations in R_{eff} and L_{eff} would have to be considered. Further, in a number of practical cases it is permissible to calculate R_{eff} and use L_2 when finding the value for Q, since

the reflected reactance may not be very large. It is often sufficient to calculate selectivity at, say, the mid-frequency of the tuning range, although it should be noted that selectivity is a function of frequency and dependent on the circuit operating conditions.

It is now worth while considering three special cases for the aerial coupling transformer :

(1) Where the aerial can be represented by a capacitive reactance in series with a resistance.

(2) Where the aerial behaves like an inductive reactance in series with a resistance.

(3) Where the aerial is coupled to a transmission line which " looks " like a pure resistance at the aerial terminals ; or the aerial itself, coupled to the receiver, looks like a pure resistance.

Any aerial can be made to appear resistive if the reactance components are tuned out by a suitable arrangement of the aerial coupling network ; this is a common procedure with fixed frequency installations.

In **case** (1) it is usual to make the primary circuit resonate below the lowest tuning frequency, since this offers a good compromise with regard to constancy of gain, tracking, etc. This arrangement is also advantageous because the effect of the aerial is merely to move the primary resonant frequency still further away from the tuning range, thereby minimizing the effect of the aerial on the secondary without losing too much gain. In general the high impedance primary winding is resonated at 0.6 to 0.8 of the lowest tuning frequency, either with the aerial capacitance alone, or with an added capacitor. In either case the resonant frequency must not be close to the intermediate frequency in a superheterodyne receiver, and normally a broadcast band primary is resonated at a lower frequency than the i-f so that no aerial which is likely to be used will resonate the aerial primary at the intermediate frequency.

Since the gain obtained from an aerial transformer with its primary resonant outside the low frequency end of the tuning range decreases towards the high frequency end of the range, a small capacitance is often added across the top of the transformer. This capacitance increases the gain at higher frequencies and allows a substantially flat gain and noise characteristic to be obtained from the transformer over the desired range. This type of aerial transformer is fairly representative of those likely to be encountered on the medium and long wavebands. Practical values for the coefficient of coupling between primary and secondary are a compromise between various factors such as gain, tracking error and signal-to-noise ratios ; usual values are from 0.15 to 0.3, with most aerial coils using values of about 0.2 (this should include any added top capacitance coupling). The value for primary Q is of the order of 50, and the secondary Q is about 90 to 130 in typical aerial coils using the range of coupling coefficients stated, although Q values greater than 200 can readily be obtained by using special iron cores. Values of secondary Q above 100 generally call for smaller values of k than 0.2 if good tracking is to be achieved.

If for some particular reason it is necessary to use a higher coefficient of coupling than say 0.4, then the primary inductance should be small and arranged to resonate with the aerial at about $\sqrt{2}$ of the highest signal frequency. This improves the signal-to-noise ratio but may give rise to tracking difficulties.

Case (2) normally applies with untuned loop aerials but the condition can occur with some types of aerials on the shortwave bands. In this case it is usual to make the coefficient of coupling fairly high, of the order of 0.5 or more, and a useful rule is to make the value of primary inductance approximately equal to 0.4 of the loop (or aerial) inductance (see Refs. B17 and A2, p. 44).

For use with domestic type receivers operating on the short wave band, the standard dummy antenna is taken as being 400 ohms resistive, even though the aerial reactance will not be zero across the tuning range, as this gives an indication of average operating conditions. The aerial coupling transformer for this type of receiver can be designed using the procedure given in Case (3) below.

The untuned loop antenna (which should be balanced to ground for best results) finds application with portable receivers, and in some cases with ordinary A-M broadcast receivers. Loops have also been used in F-M receivers operating on the 40-50

Mc/s band, but not in receivers for the 88-108 Mc/s band because of the ease with which a half-wave dipole can be arranged inside the receiver cabinet.   With loops it is necessary to keep stray capacitances to a minimum so as to avoid resonance effects within or close to the band of frequencies to be received ; this also assists in achieving good tracking because a loop that behaves like a small fixed inductance can be compensated for quite readily in the aerial coil secondary.

The number of turns used in untuned loop aerials is not very critical, since a loss in effective height by using fewer turns can be offset by the increased voltage step-up possible with the coupling transformer.   To keep stray capacitances low the fewer turns used the better.   The best compromise for receivers operating on the medium or short-wave bands is a single turn loop, or perhaps a few turns (say not more than four) of wire wrapped around the carrying case for medium wave portable receivers (this wire can be litz or copper braid, and even rubber covered hook-up wire is often satisfactory).   The best type of single turn loop is one made out of copper tubing (or any other good conductor such as aluminium or one of its alloys) with the tube diameter not less than about 0.25 in. and preferably considerably larger than this ; the loop area should always be as large as possible.   If the loop is to be used on the long-wave band it is preferable to use a few turns (about ten) of litz wire rather than a single turn loop.   This is necessary because of the drop in Q which occurs with the single turn loop at the lower frequencies.   However, satisfactory results have been achieved with a 0.25 in. diameter copper tube loop at a frequency of 500 Kc/s.

In many cases a two winding transformer is not used and the loop is made part of the tuned circuit (being in series with the main tuning inductance and earthed at one side).   An alternative arrangement is to tap the loop across part of the main tuning inductance ; this can be made to give very satisfactory results.

Because the inductance of loops (of the type being considered) is usually very small, it is often difficult to measure the Q at the working frequencies.   This difficulty arises with standard types of Q meters because the maximum capacitance is limited to about 400 $\mu\mu$F.   A useful method (suggested by J. B. Rudd) is as follows :

Couple the loop fairly tightly to an inductance which is sufficiently large to be tuned to the working frequency.   Measure the Q of the inductance with the loop open circuited (call this Q_1) and then the Q (call this Q') with the loop short circuited at its terminals.   The Q of the loop (Q_L) is given by

$$Q_L = \frac{yQ_1Q'}{Q_1(1 - y) - Q'}$$

where Q_1 = magnification factor for the primary inductance with the loop open circuited.

Q' = magnification factor for the primary inductance with the loop terminals short circuited.

$y = 1 - (L'/L_1)$

L' = inductance of primary with loop short circuited.

and L_1 = inductance of primary with loop open circuited.

See (vi) below for some discussion of directly tuned loops.

Case (3) permits of a fairly simple solution.   Normally k is about 0.2 or less, and the primary winding reactance (at the geometrical mean frequency of the tuning range) is made equal to the characteristic impedance of the transmission line or aerial.   A particularly simple solution has been made by Rudd (Ref. B12) to give conditions for two point matching in the tuning range.   The procedure is as follows :

Geometrical mean angular frequency $= \omega_0 = \sqrt{\omega_1 \omega_2}$ where ω_1 and ω_2 are the lower and upper angular frequency band limits.   The primary inductance $= L_1 = Z_0/\omega_0$ where Z_0 is the characteristic impedance of the transmission line.

The required coefficient of coupling is then determined from

$$k = \sqrt{\frac{1}{Q_s}\left[\frac{1 + \alpha}{\alpha^{\frac{1}{2}}}\right]} \tag{8}$$

where $\alpha = \omega_2/\omega_1 = f_2/f_1$

and Q_s = secondary circuit Q in the absence of the primary. The value for Q_s should include the effect of valve loading, and so

$$Q_s = \frac{QR}{Q\omega_0L_2 + R} \tag{9}$$

where Q is for the unloaded secondary
and R is the valve input resistance.

A useful approximate formula for estimating the aerial coil gain under this condition is

$$\frac{E_2}{E_1} = \frac{1}{2}\sqrt{\frac{Q_s\omega L_2}{Z_0}}. \tag{10}$$

This expression gives the maximum possible value of gain and corresponds to eqn. (7a) given previously.

Although elaborate methods are available for calculating optimum signal-to-noise ratios etc., the details above should give a fairly close approximation to the required practical conditions where the aerial coupling circuit is a compromise between the various conflicting factors discussed previously. These factors of gain, selectivity etc. should be carefully considered as in nearly every case the performance of a receiver is largely governed by the aerial coupling circuit.

(iv) Tapped inductance

Another common type of aerial coupling arrangement is shown in Fig. 23.3. This arrangement has found wide application in receivers tuning the 88-108 Mc/s F-M band because of its simplicity and the ease with which the correct aerial tapping point may be found. This can also be considered as a particular case of the generalized aerial circuit. This circuit can be treated in a simplified manner due to Zepler (Ref. A2). The required tapping point on the total coil, which is usually a solenoid, can be found approximately from the turns ratio

FIG. 23-3 TAPPED INDUCTANCE AERIAL TRANSFORMER

$$\frac{n_2}{n_1} = \sqrt{\frac{Q_s\omega L_2}{Z_0}} \tag{11}$$

where n_2 = total number of turns
  n_1 = number of turns across which aerial is connected
  ω = 2π × operating frequency
  Q_s = magnification factor for complete circuit
and Z_0 = impedance of the aerial (or the characteristic impedance of a transmission line).

The gain is found from equation (10),

$$\frac{E_2}{E_1} = \frac{1}{2}\sqrt{\frac{Q_s\omega L_2}{Z_0}}$$

The selectivity is calculated using the same methods as before, from the equivalent circuit. Whether this equivalent circuit takes the form of a series or parallel tuned circuit is usually not important, as the results obtained are substantially the same, and provided the circuit Q exceeds about 10 the Universal Selectivity Curves can be applied directly in either case.

It should be noted that the treatment is only approximate, as the exact tapping point is dependent on the coupling between the two sections of the aerial coil winding. However, for a wide variation in coupling, the correct tapping point is not very different from that obtained by the above procedure, which assumes unity coupling, and the loss in secondary voltage as compared with optimum coupling is quite small.
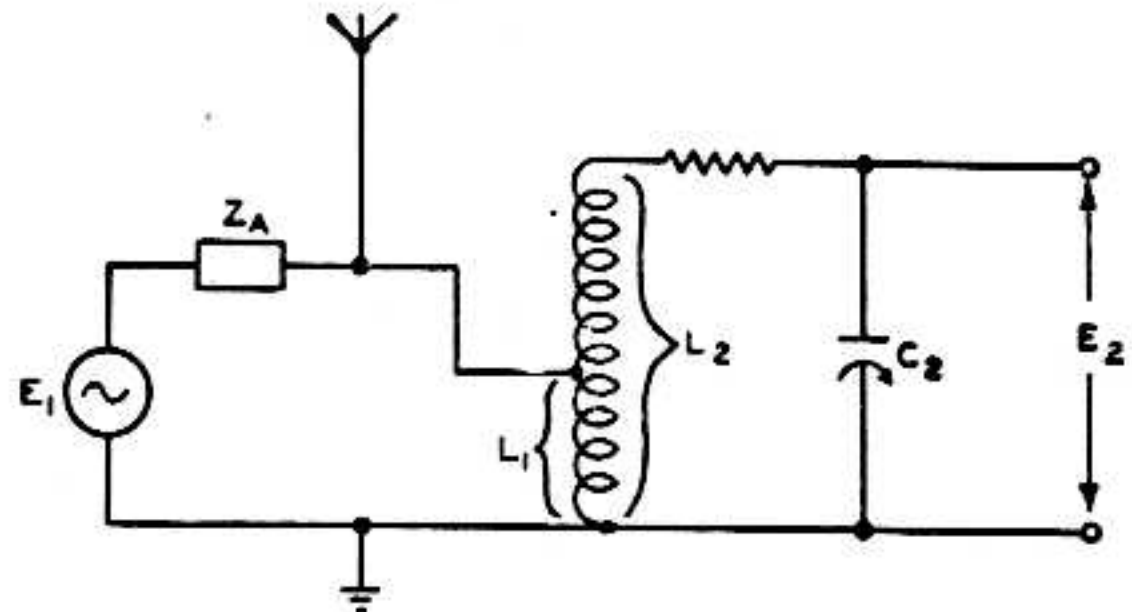
Valve input loading is usually a problem at v-h-f and a procedure to minimize this effect is to tap down across L_2. This will allow a higher value of dynamic impedance for the tuned circuit with improved selectivity, and under some conditions the increase in impedance can more than offset the loss in voltage gain caused by tapping down.

(v) Capacitance coupling

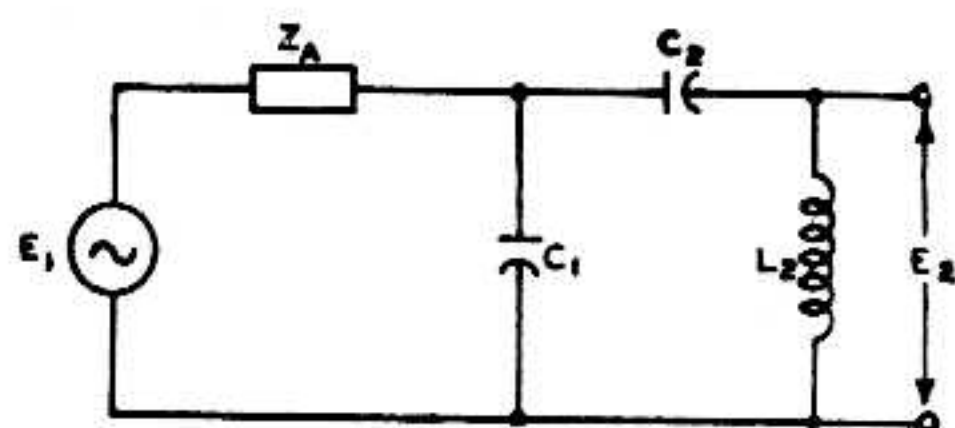A further type of coupling arrangement fairly commonly used is shown in Fig. 23.4.

FIG. 23.4

Fig. 23.4. *Capacitance Coupled Aerial Transformer.*

This shows a common disadvantage with the circuit of Fig. 23.3 in that the variation in gain across the tuning range is quite appreciable. The treatment is quite straightforward and is left to the reader ; a fairly complete analysis is given in the references.

(vi) General summary

From the considerations detailed above, it can be seen that the coupled circuit arrangement of Fig. 23.2 is generally most satisfactory since it readily lends itself, with minor modifications, to applications using balanced or unbalanced aerial systems. Under some conditions of receiver operation other methods are used because of the practical consideration of ease of adjustment for best operation, or for use with a particular type of aerial system.

Fixed tuned aerial stages are sometimes used with receivers covering a limited range of frequencies. A typical case would be in the range of 88-108 Mc/s where the loaded Q of the aerial coil may be less than 20. In such a case a simple arrangement is for the secondary to be resonated near the centre frequency in the band ; this introduces gain variations of about 2 : 1 across the tuning range. The gain variations can be offset by resonating the secondary at a point in the band such that the gain variations offset those in r-f or mixer stages. The overall gain variations for the band under consideration can be limited to less than 1.5 : 1 with a superheterodyne receiver using a tuned r-f stage, and as the loss in gain and signal-to-noise ratio is not very serious, the saving in the cost of a gang section is often well worth while when weighted against the other factors. If the aerial circuit loaded Q is above about 20, this method is not recommended because of the deterioration in circuit performance. In some cases it is preferable to tune the aerial circuit and use an untuned r-f stage (when this is practicable).

One type of coupling circuit which has not been mentioned previously is the directly tuned loop. In this case the loop provides the tuned circuit inductance as well as the direct signal pick-up. As for any other loop, the area and the Q should be as large as possible. This loop will probably be used, in most cases, with portable receivers. If the loop can be placed in such a position that its Q is not materially affected by the presence of the receiver and batteries, then its performance will usually be superior to that of the single turn loop (an improvement of about 6 db can be expected). However, this arrangement is seldom convenient and in many cases better results are possible with the single turn loop (or one with a few turns as discussed previously) because any effect on its Q can be offset to a large extent by the use of an aerial coupling transformer having a high Q winding in the tuned circuit. The improvement can be seen most readily by considering a low Q loop, having a small inductance, connected directly in series with an aerial coil which has a high Q and whose inductance is practically the whole of the tuned circuit inductance.

In some cases it is required to use an additional external aerial with the loop to increase signal pick-up. There are many possible arrangements but amongst the simplest is the use of a small capacitance (say 10 to 20 $\mu\mu$F) connected between the

external aerial and the top of the tuned circuit (i.e. to the grid connection) or alternatively a tap on the aerial coil secondary a few turns up from the earthy end. For a directly tuned loop the aerial is tapped into the loop in the same way as for the other coil arrangement ; alternatively the series capacitor arrangement is satisfactory. The main disadvantage of both these simple arrangements is the large change in gain which occurs across the tuning range, but for medium-wave commercial portable receivers additional circuit complications are seldom justified.

The subject of aerial-to-receiver coupling is a large one and only a few of the more important design factors have been considered. For more complete information on particular systems it is necessary to consult the text books and references.

SECTION 3 : R-F AMPLIFIERS

(i) Reasons for using r-f stage (ii) Mutual-inductance-coupled stage (iii) Parallel tuned circuit (iv) Choke-capacitance coupling (v) Untuned and pre-tuned stages (vi) Grounded grid stages.

(i) Reasons for using r-f stage

The necessity for adding a r-f amplifier in a superheterodyne receiver arises from the need for

(1) greater gain
(2) improved rejection against undesired signals
(3) improved image frequency rejection
(4) reduction of the effects of spurious frequency combinations,
(5) improved signal-to-noise ratio,
(6) the prevention of radiation at the local oscillator frequency (under some conditions).

In the case of the t-r-f receiver, additional r-f stages are usually added for reasons (1) and (2) above.

The design of a r-f stage is quite similar for both types of receiver and so they will not be considered separately. The only point which will be mentioned is that, in the case of the t-r-f receiver the final r-f amplifier feeds into a detector stage which, if a diode is used, heavily loads the tuned circuit unless a tapping point is used. For this reason it is common practice to incorporate anode bend or linear reflex detectors which do not appreciably load the input circuit. However, a diode is convenient as a means of supplying an a.v.c. bias voltage of the correct polarity, a condition not fulfilled by the other two types of detectors mentioned without additional circuit arrangements.

(ii) Mutual-inductance-coupled stage

The most common arrangement for coupling a r-f amplifier valve to the following stage (Fig. 23.5) is by means of a transformer using a tuned secondary with the primary winding resonated either above or below the tuning range. The primary winding resonates with the valve output capacitance, winding capacitance, stray wiring capacitances and sometimes an added capacitor, these being lumped together and shown as C_p. If the winding is of the low impedance type it is made to resonate at approximately 1.2 to 1.5 times the highest tuning frequency. Under this condition the load presented to the valve is inductive and thus will give rise to regeneration by introducing a negative resistance component plus a capacitance into the r-f valve input circuit.
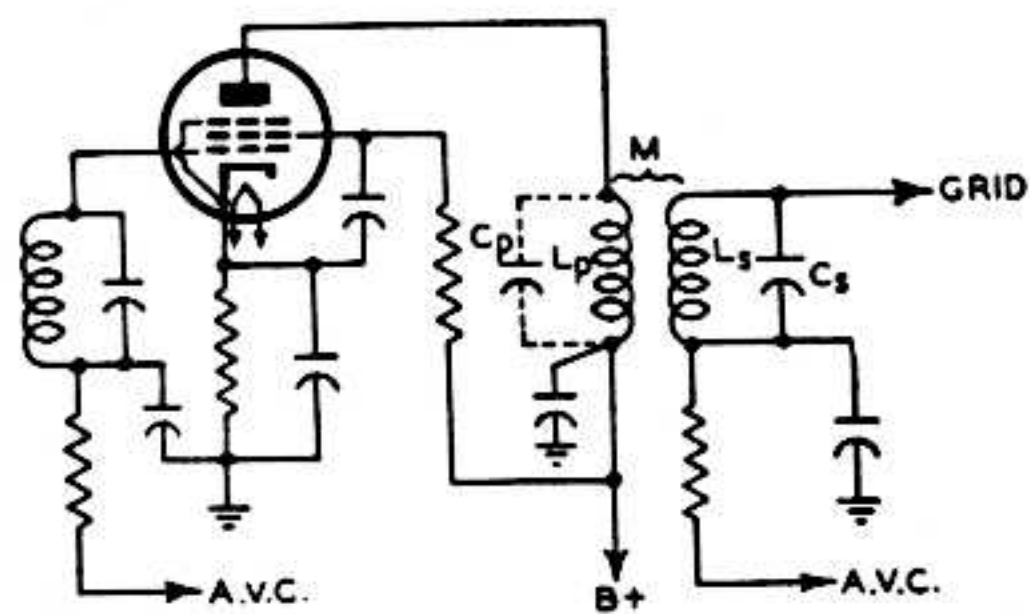

FIG.23-5 TRANSFORMER COUPLED R-F STAGE.

If the winding is of the high impedance type, then it is usual to resonate the primary at approximately 0.6 to 0.8 of the lowest tuning frequency, but the resonant frequency should not be close to the intermediate frequency in superheterodyne receivers. Under this condition the valve has a capacitive load which causes degeneration by introducing a positive resistance component plus a capacitance into the r-f valve input circuit due to the " Miller Effect." It is not usual to resonate the primary winding within the tuning range because of the very large changes in gain and selectivity which would be introduced. In addition, the problem of tracking the various tuned circuits becomes almost hopeless.

When the aerial coil primary is of the low impedance type it is usual, although not essential, for the r-f stage to be of this type also and vice-versa. This simplifies tracking problems.

The methods to be used for calculating the gain and selectivity of this arrangement are set out in detail in Chapter 9 Sect. 6(ii) under the heading " Coupled Circuits—Tuned Secondary." For the usual practical arrangement the stage gain is given approximately by $g_m Q_2 \omega M$,

where g_m = mutual conductance of r-f amplifier valve,

ω = $2\pi \times$ operating frequency,

M = $k\sqrt{L_1 L_2}$ = mutual inductance,

and Q_2 = secondary circuit Q including effect of valve input resistance, but not " reflected " resistance from the primary.

The selectivity can also be calculated, approximately, by the method suggested for aerial coils.

When designing a stage of this type it is necessary, as in the case of the aerial coil, to arrive at a compromise between gain, selectivity and tracking errors. This usually leads to a value for the coefficient of coupling of approximately 0.15 to 0.3, with the lower values preferred for good tracking between aerial, r-f, and, in the case of the superhet., the oscillator stage.

As with aerial coils, the primary Q is about 50 (or perhaps less) and the secondary Q from say 90 to 130 in typical coils using this range of k values. Higher Q generally calls for the lower values of k.

In the case of receivers working at say 50-400 Kc/s it is often inconvenient to resonate the primary winding in the manner suggested. Under these circumstances it is sometimes necessary to damp the primary winding heavily so that no pronounced resonance occurs. It is then permissible to resonate the winding within the tuning range. This procedure can lead to a severe loss in gain but is sometimes justified under practical conditions where the loss of gain is outweighed by other factors.

The addition of top capacitance coupling, i.e. a small capacitance from the plate connection to the grid connection of the following stage, is often employed to equalize gain variations across the tuning range. The value of the capacitance is usually of the order of a few micro-micro-farads and the exact value is best determined experimentally. It should be noted that the presence of this capacitance will alter the coefficient of coupling, stage gain, selectivity and the primary resonant frequency, as well as circuit tracking. There is always some capacitance coupling present with any practical transformer and this largely determines the relative physical arrangement of the primary and secondary windings, which are connected (in nearly all practical cases) so that the capacitance coupling adds to the mutual inductance coupling.

(iii) Parallel tuned circuit

A simple parallel tuned circuit is sometimes used in the r-f stage, and can be made to give higher gains than the transformer coupled arrangement. The difficulties encountered are that

(1) large gain variations occur across a band of frequencies
(2) the skirt selectivity is rather poor
(3) tracking with conventional aerial circuits is difficult
(4) the tuning capacitor has to be isolated from B+.

For these and other reasons this circuit is not in common use in broadcast receivers, but it might be incorporated in a receiver working at high frequencies covering a

restricted tuning range with some possible advantage. The gain of the stage is given by $g_m Q_{eff} \omega L$

where g_m = mutual conductance of r-f amplifier valve,

$\quad\quad Q_{eff}$ = Q of r-f coil when loaded by plate resistance of r-f amplifier and input resistance of following stage,

$\quad\quad \omega$ = $2\pi \times$ frequency,

and $\quad L$ = inductance of tuned circuit.

Selectivity is calculated as for any other single tuned stage, using the methods of Chapter 9 or Chapter 26, Sect. 4(iv)A.

(iv) Choke-capacitance coupling

Fig. 23.6 shows a common r-f coupling circuit for use at v-h-f such as on the 88-108 Mc/s F-M broadcast band. The usual arrangement takes the form of a r-f choke, resonated well below the lowest tuning frequency, coupled to a tuned circuit $L_2 C_2$ by means of a capacitor C. The choice of a suitable value for C will allow some reduction in variations of stage gain across the tuning range as it can be considered as part of a voltage divider formed with the tuned grid circuit. A further useful function of the arrangement is that when a suitable resonance frequency has been chosen for the choke, the total stray capacitance across the tuned circuit is effectively reduced, thereby reducing limitations on the tuning range ; a small value for C assists in this latter regard. However, the application of the circuit is largely one of practical convenience and, as always, is a compromise.
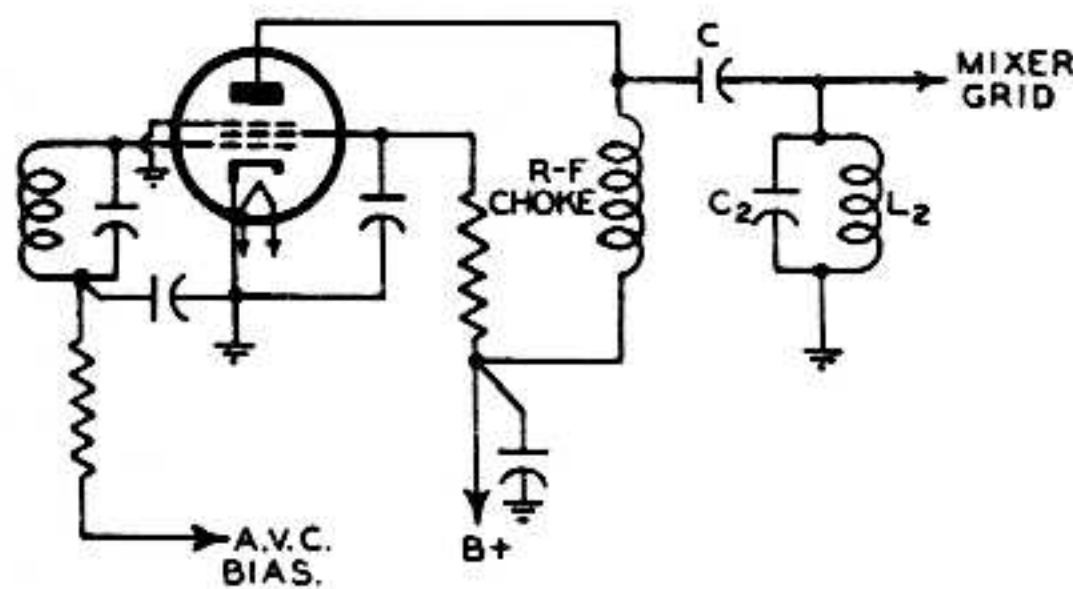

FIG. 23·6 CAPACITANCE COUPLED R-F STAGE.

With a choke resonated at $1/\sqrt{2}$ of the lowest tuning frequency in a receiver with a low frequency limit of 87.5 Mc/s and an upper limit of 108.5 Mc/s, the values obtained for the additional capacitance shunted across the tuned circuit were 3.6 $\mu\mu$F at 88 Mc/s and 4 $\mu\mu$F at 108 Mc/s, excluding the input capacitance of the following stage. The value of C was 10 $\mu\mu$F.

Because of the low value of dynamic impedance for this circuit at v-h-f, the effect of the plate resistance of the r-f valve can usually be neglected (but see Ref. A8). In this case the selectivity for the circuit can readily be calculated as for a single tuned circuit, including the effects of the input resistance of the following stage, without introducing very large errors. The gain may be calculated with sufficient accuracy as for the case of the simple tuned circuit acting as the anode load and multiplying the result by a suitable factor to allow for the voltage division due to the capacitive reactances at the frequency being considered. This factor is readily found from a knowledge of the equivalent capacitance due to the choke circuit, at the working frequency, and the value selected for C.

(v) Untuned and pre-tuned stages

Untuned, or in some cases pre-tuned, r-f stages are sometimes used in receivers where economy and/or limited space are important factors, or where only a small frequency range is to be covered. Unfortunately in the case of portable battery receivers, where the application could be valuable, the use of untuned r-f stages is made difficult by the fairly low values of mutual conductance for the usual r-f amplifier valves. The gain obtained is limited and it may be preferable to use an additional i-f stage if the r-f stage cannot be tuned and gain is a prime requirement. With mains operated receivers the problem is different, as many suitable valve types are available and quite good stage gains (about 5 times or more, on the medium and short wave bands) can be obtained over a wide range of frequencies. Suitable design methods using untuned stages for the medium and short wave bands have been given in R.C.A. Application Note No. 116 (reprinted in Radiotronics No. 124) to which the reader is referred for full details A typical coupling circuit of the type discussed in

the Application Note is shown in Fig. 23.7.

FIG. 23.7 UNTUNED R-F STAGE.

(vi) Grounded-grid stages

As previously mentioned this type of amplifier will not be discussed in detail. However, the following equations should prove to be of assistance, and are generally a sufficiently good indication since experimental techniques usually provide the most suitable method for determining the best operating conditions for this type of amplifier.

$$\text{Voltage gain} = \frac{R_L}{r_p + R_L}\,(\mu + 1)$$

$$\text{Resistance loading across input circuit} \approx \frac{r_p + R_L}{\mu + 1}$$

(loading across output circuit $\approx r_p$).

For $\mu \gg 1$ the optimum load impedance is given by

$$R_L = r_p\sqrt{1 + R_i g_m}$$

$$\text{Noise factor} = N = \sqrt{2 + \frac{R_1}{R_0 A_1{}^2} + \frac{R_2}{R_0 A_1{}^2 A_2{}^2}}$$

where R_L = load impedance connected between plate and B+
  r_p = valve plate resistance
  μ = valve amplification factor
  g_m = valve mutual conductance
  R_i = valve input resistance at working frequency
  R_1 = equivalent noise resistance of first valve
  R_2 = equivalent noise resistance of second valve
  R_0 = generator impedance e.g. aerial resistance
  A_1 = voltage gain of input circuit
and A_2 = voltage gain of first stage, excluding the voltage gain of the input circuit.

It should be noted that it is always advisable to operate the heater and cathode of the valve at the same r-f potential. This is readily achieved by using suitable chokes in the heater leads. For more detailed information on grounded-grid amplifiers see References B8, B18, A10 and A4.
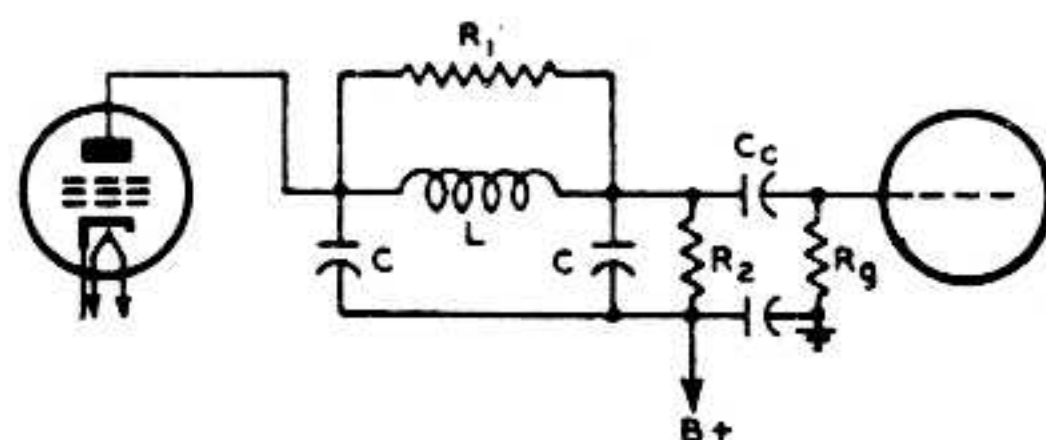
SECTION 4 : IMAGE REJECTION

(i) *Meaning of image rejection* (ii) *Image rejection due to aerial stage* (iii) *Other considerations.*

(i) Meaning of image rejection

With the superheterodyne receiver it is necessary to consider not only adjacent channel selectivity due to the aerial and r-f stages but also whether the selectivity of these stages is sufficient to prevent signals on image frequencies from passing into the i-f amplifier. If there were no tuned circuits ahead of the converter valve a signal lower in frequency than the oscillator frequency, and which differs from the oscillator frequency by the value of the intermediate frequency, would pass into the i-f amplifier. Similarly a signal higher in frequency than the oscillator frequency, and which differs from the oscillator frequency by the value of the intermediate frequency, would also pass into the i-f amplifier.

When one of these frequencies is the desired one (it may be either the lower or the higher), the other is referred to as its image frequency. Clearly the desired and undesired signal frequencies are separated by twice the intermediate frequency. Putting this statement in symbolic form :

$$f_{image} = f_{osc} + f_{i-f} \text{ when } f_{osc} > f_{sig}$$
$$f_{image} = f_{osc} - f_{i-f} \text{ when } f_{osc} < f_{sig}$$

where f_{sig} = desired signal frequency.

As an example : On the 540-1600 Kc/s A-M broadcast band the oscillator frequency is normally above the signal frequency. If the reasons for this are not obvious, consideration of the oscillator tuning range and the values of variable capacitance required will make it so. The usual i-f is 455 Kc/s and if a signal frequency of 1000 Kc/s is taken,

then f_{image} = (1000 + 455) + 455 = 1910 Kc/s.

Since the converter cannot discriminate between the two signals of 1000 Kc/s and 1910 Kc/s, it is necessary for the aerial circuit, and r-f circuit if this is used, to make the magnitude of the 1000 Kc/s voltage much greater than the magnitude of the 1910 Kc/s voltage. Otherwise severe interference can result when signals are being transmitted at both these frequencies.

When making calculations for image rejection the worst conditions are usually found at the high frequency end of the tuning range, provided that the selectivity of the tuned circuits does not vary appreciably with change in frequency. This is because the separation between the two frequencies is a smaller proportion of the frequency to which the receiver is tuned. It should also be clear that higher values of i-f will materially assist in reducing image effects because of the wider frequency separation between the desired and undesired signals ; it is for this reason that some v-h-f receivers use intermediate frequencies of the order of 10.7 Mc/s.

(ii) Image rejection due to aerial stage

The image rejection due to an aerial stage alone is calculated from

$$\sqrt{1 + Q^2\left(\frac{f}{f_0} - \frac{f_0}{f}\right)^2}$$

where $Q = \omega L_{eff}/R_{eff}$ (i.e. loaded value for secondary circuit),

$L_{eff} \approx L_2$, the secondary circuit inductance,

R_{eff} = total secondary circuit series resistance including that due to reflection from primary circuit and valve input resistance (Valve input admittance is discussed in detail in Chapter 2 Sect. 8, and further comments are given in Sect. 5 of this chapter),

f = image frequency when oscillator frequency is higher than signal frequency,

and f_0 = signal frequency under oscillator conditions as for f.

If the oscillator is lower in frequency than the signal frequency then f and f_0 should be interchanged.

Exactly the same expression and procedure are used to calculate the image protection afforded by a r-f stage. Since the calculated image rejection is a voltage ratio, the ratios obtained for the various stages are multiplied together, or the ratios expressed in decibels can be added to give the total image protection provided by the input circuits to the converter.

(iii) Other considerations

The image protection due to the r-f stage may be greater than that due to the aerial stage. A typical example occurs when the loading on the aerial stage due to the r-f amplifier valve has a positive value of input resistance and the loading on the r-f stage due to a pentagrid converter, using inner grid oscillator voltage injection, has a negative value of input resistance. The resultant Q, simply from these considerations, may then be higher in the case of the r-f coil. Also, since the resultant Q of the aerial

coil is halved, under some conditions of operation, by using two point matching to an aerial transmission line, the greater part of the image protection may here again be due to the r-f stage. These considerations are particularly important at the higher frequencies. On the long and medium wave bands sufficient image protection can often be obtained from the discrimination afforded by the aerial coil alone, but on the shortwave bands deterioration in image rejection is serious (and rapid) unless the intermediate frequency is increased above the usual value of about 455 Kc/s.

It is important to note that a good image rejection ratio is desirable since this also indicates the selectivity of the signal circuits and the degree of rejection against spurious frequency combinations. A high degree of selectivity preceding the first r-f amplifier valve has obvious advantages in reducing cross-modulation effects.

SECTION 5 : EFFECTS OF VALVE INPUT ADMITTANCE

(i) Important general considerations (ii) Input loading of receiving valves at radio frequencies (A) Input conductance (B) Cold input conductance (C) Hot input conductance (D) Change in input capacitance (E) Reduction of detuning effect.

(i) Important general considerations

A fairly complete discussion of some of the factors relating to valve input admittance has been given in Chapter 2 Sect. 8. These considerations are additional to the following details which take into account electron transit time effects at frequencies up to about 100-150 Mc/s, where conventional receiver design techniques begin to fail.

From the receiver designer's point of view the main factors (if we neglect changes in valve output admittance—but see Ref. A8) are

(1) A knowledge of the actual values of input resistance and capacitance (neither of which is constant with changes in valve operating conditions) with which he will have to contend when these are shunted across the tuned circuit (this must include the effects of feedback etc.), and

(2) A knowledge of the effects on input capacitance, in particular, when the signal voltage changes, giving rise to an alteration in a.v.c. bias. From this information the amount of detuning of the circuits can be determined. As the input components are in parallel with the tuned circuits, it is often more convenient to discuss the values of conductance, susceptance, and admittance rather than resistance, reactance, and impedance. Several extremely useful lists of values of conductance and capacitance for some common types of receiving valves are given in Chapter 2 Sect. 8, and further details of some additional types are set out later in this section. In cases where the value of short circuit input conductance at a required frequency is not given, then a close approximation can be found by multiplying the conductance value by $\left(\dfrac{\text{Frequency required}}{\text{Frequency stated}}\right)^2$; but extrapolation should not be carried too far e.g. a conductance value at 100 Mc/s should not be used to find a conductance at 455 Kc/s, but will give a reasonable approximation at, say, 10 Mc/s.

A **figure of merit,** which forms a useful basis for comparing various types of r-f voltage amplifier valves, is given by

$$\frac{g_m}{\sqrt{g_i}} \text{ (or } g_m\sqrt{R_i}\text{)}$$

where g_m = control grid to plate transconductance (mutual conductance)
  g_i = short circuit grid input conductance
and R_i = short circuit grid input resistance.

At the higher frequencies cathode lead inductance is important.

If the valve types on which information regarding short circuit input admittance is required are not listed, experimental techniques can always be resorted to and reason-

ably accurate results are possible using a high frequency Q meter to determine the capacitive and resistive loading effects. The usual precautions as to length of leads, earthing etc. must be carefully observed when making these measurements, if the results are to be of any practical value. Something of this nature is usually required when considering the effects of the converter valve on its input circuit, since the published information is rather meagre. In this regard it is well to remember that some types of converter valves, such as those using inner grid injection, e.g. types 6A8, 6K8, 6SA7, 6BE6 etc., give negative loading whilst other types such as X61M, X79, 6J8-G, 6L7, etc. (which use outer grid oscillator injection) give positive loading. Increasing the negative bias on the signal grid will reduce the loading effect with all types, but at very high values of bias the negative loading may reverse its polarity and become positive. [The reasons for these effects are discussed later in this section in connection with input loading of receiving valves.] To reduce input conductance some valve types have more than one cathode connection (see Ref. A8), and typical examples are types 6AK5 and 6AG5, each having two cathode terminals. Even when the cathode terminals are directly connected to ground the length of the cathode lead is still sufficient to provide appreciable inductive reactance at frequencies of the order of 100 Mc/s. For this reason all plate and screen by-pass capacitors should be returned to one cathode lead and the grid returns to the other. The alternating voltages developed across the cathode lead inductance due to currents from the plate and screen circuits are in this way prevented from being directly impressed in series with the grid circuit, since it is the grid to cathode voltages which are of major importance. Direct current divides between the two available cathode paths, but this is not important since the path taken by this current normally does not affect the grid input admittance. To obtain grid bias the " grid " cathode lead employs the usual resistor and capacitor combination. The r-f by-pass capacitors from the plate and screen circuits connect directly to the other cathode terminal which is not directly connected to ground in this arrangement. A typical circuit is shown in Fig. 23.8.

In pentodes working at v-h-f it is advisable to connect the suppressor grid terminal directly to ground, when a separate terminal is available for this electrode, rather than to the cathode, because the suppressor lead inductance would then be connected in series with any external cathode lead inductance. If this precaution is not taken the coupling between control grid and plate is increased, because of the capacitance from control grid to suppressor and from suppressor to plate, the junction of these two

FIG. 23·8 R-F AMPLIFIER WITH TWO CATHODE CONNECTIONS.

capacitances having an impedance to ground depending on the total lead inductance.

Similar considerations also apply to screen grid circuits, particularly when the valve has a single cathode lead, and the shielding action of this grid can be seriously affected if proper precautions are not taken, such as directly earthing by-pass capacitors. These circuit arrangements are not very important at low frequencies where degenerative effects may be more serious, necessitating the connection of the suppressor grid and by-pass capacitors directly to the cathode.

(ii) Input loading of receiving valves at radio frequencies*

(See also Chapter 2, Sect. 8 and R.C.A. Application Note No. 118—Ref. B25).

The input resistance of r-f amplifier valves may become low enough at high radio frequencies to have appreciable effect on the gain and selectivity of a preceding stage. Also, the input capacitance of a valve may change enough with change in a.v.c. bias to cause appreciable detuning of the grid circuit. It is the purpose of this Note to discuss these two effects and to show how the change in input capacitance can be reduced.
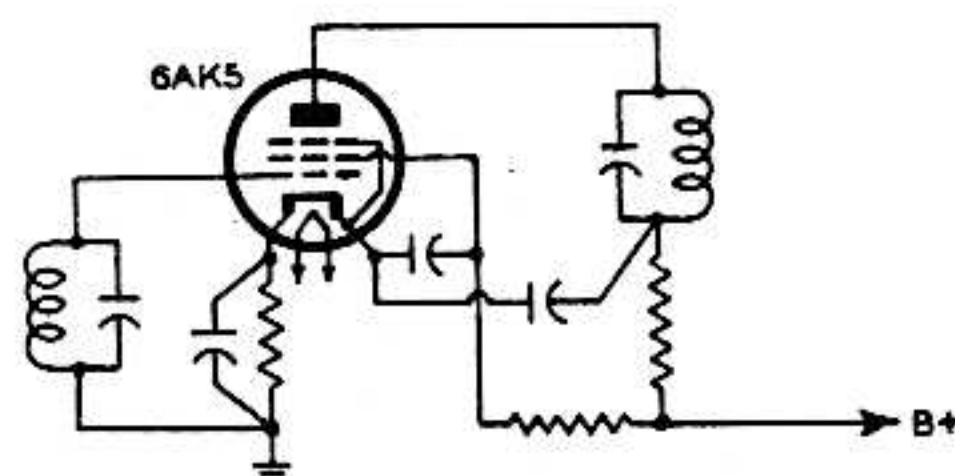
*From R.C.A. Application Note No. 101.

(A) Input conductance

It is convenient to discuss the input loading of a valve in terms of the valve's input conductance rather than input resistance. The input conductance g_i of commercial receiving valves can be represented approximately by the equation

$$g_i = k_c f + k_h f^2 \tag{1}$$

where f is the frequency of the input voltage. A table of values of k_c and k_h for several r-f valve types is shown below. The approximate value of a valve's input conductance in micromhos at all frequencies up to those in the order of 100 megacycles can be obtained by substituting in Eq. (1) values of k_c and k_h from the table. In some cases, input conductance can be computed for conditions other than those specified in the table. For example, when all the electrode voltages are changed by a factor n, k_h changes by a factor which is approximately $n^{-1/2}$. The value of k_c is

TABLE OF APPROXIMATE VALUES OF k_c AND k_h FOR SEVERAL VALVE TYPES

Valve Type	Description	Plate volts	Screen volts	Signal-Grid bias volts	Supressor volts	k_c micro-mhos/ Mc/s	k_h micro-mhos/ Mc/s^2
6A8	Pentagrid	250	100	−3	—	0.3	−0.05*
6J7	Pentode	250	100	−3	0	0.3	0.05
6K7	Pentode	250	100	−3	0	0.3	0.05
6K8	Triode-hexode	250	100	−3	—	0.3	−0.08†
6L7	Mixer	250	100	−3	—	0.3	0.15‡
6SA7^1¶	Pentagrid	250	100	0	—	0.3	−0.03§
6SA7^2¶	Pentagrid	250	100	−2	—	0.3	−0.03§
6SJ7¶	Pentode	250	100	−3	0	0.3	0.05
6SK7¶	Pentode	250	100	−3	0	0.3	0.05
954	Pentode	250	100	−3	0	0.3	0.005
1851 6AC7/	Pentode	250	150	−2	0	0.3	0.13
1852¶ 6AB7/	Pentode	250	150	−2	0	0.3	0.13
1853¶	Pentode	250	200	−3	0	0.3	0.065

*For oscillator-grid current of 0.3 mA through 50 000 ohms.
†For oscillator-grid current of 0.15 mA through 50 000 ohms.
‡For wide range of oscillator currents.
§For grid No. 1 current of 0.5 mA through 20 000 ohms.
^{1}Self-excited. ^{2}Separately excited.
¶Denotes single ended valve type.

practically constant for all operating conditions. Also, when the transconductance of a valve is changed by a change in signal-grid bias, k_h varies directly with transconductance over a wide range. In the case of converter types, the value of k_h depends on oscillator-grid bias and oscillator voltage amplitude. In converter and mixer types, k_h is practically independent of oscillator frequency.

In eqn. (1), the term $k_c f$ is a conductance which exists when the cathode current is zero. The term $k_h f^2$ is the additional conductance which exists when cathode current flows. These two terms can be explained by a simple analysis of the input circuit of a valve.

(B) Cold input conductance

The input impedance of a valve when there is no cathode current is referred to as the cold input impedance. The principal components of this cold impedance are a resistance due to dielectric hysteresis, and a reactance due to input capacitance and cathode-lead inductance. Because these components are in a parallel combination, it is convenient to use the terms admittance, the reciprocal of impedance, and sus-

ceptance, the reciprocal of reactance.   For most purposes, the effect of cathode-lead inductance is negligible when cathode current is very low.   The cold input admittance is, therefore, a conductance in parallel with a capacitive susceptance.   The conductance due to dielectric hysteresis increases linearly with frequency.   Hence, the cold input conductance can be written as $k_c f$, where k_c is proportional to the power factor of grid insulation and is the k_c of eqn. (1).

(C) Hot input conductance

The term $k_h f^2$, the input conductance due to the flow of electron current in a valve, has two principal components, one due to electron transit time and the other due to inductance in the cathode lead.   These two components can be analysed with the aid of Fig. 23.9.   In this circuit, C_h is the capacitance between grid and cathode when cathode current flows, C_g is the input capacitance due to capacitance between grid and all other electrodes, except cathode, g_t is the conductance due to electron transit time, and L is the cathode-lead inductance.   Inductance L represents the inductance of the lead between the cathode and its base pin, together with the effect of mutual inductances between the cathode lead and other leads near it.   Analysis of the circuit of Fig. 23.9 shows that, with L small as it generally is, the input conductance, g_h, due to the presence of cathode current in the valve, is approximately

$$g_h = g_m \omega^2 L C_h + g_t \qquad (2)$$

where $\omega = 2\pi f$.   The term $g_m \omega^2 L C_h$ is the conductance due to cathode-lead inductance.   It can be seen that this term varies with the square of the frequency. In this term, g_m is the grid-cathode transconductance because the term is concerned with the effect of cathode current flowing through L.   In a pentode, and in the 6L7, this transconductance is approximately equal to the signal-grid-to-plate transconductance multiplied by the ratio of direct cathode current to direct plate current. In the converter types 6A8, 6K8 and 6SA7, the signal-grid-to-cathode transconductance is small.   Cathode circuit impedance, therefore, has little effect on input conductance in these types.

For an explanation of the conductance, g_t, due to electron transit time, it is helpful to consider the concept of current flow to an electrode in a valve.   It is customary to consider that the electron current flows to an electrode only when electrons strike the surface of the electrode.   This concept, while valid for static conditions, fails to account for observed high-frequency phenomena.   A better concept is that, in a diode for example, plate current starts to flow as soon as electrons leave the cathode. Every electron in the space between cathode and plate of a diode induces a charge on the plate ; the magnitude of the charge induced by each electron depends on the proximity of the electron to the plate.   Because the proximity changes with electron motion, there is a current flow to the plate through the external circuit due to the motion of electrons in the space between cathode and plate.

Consider the action of a conventional space-charge-limited triode as shown in Fig. 23.10.   In this triode, the plate is positive with respect to cathode and the grid is negatively biased.   Due to the motion of electrons between cathode and grid, there is a current I_a flowing into the grid.   In addition, there is another current I_b flowing out of the grid due to the motion of electrons between grid and plate receding from the grid.   When no alternating voltage is applied to the grid, I_a and I_b are equal and the net grid current I_g is zero.

Suppose, now, that a small alternating voltage (e_g) is applied to the grid.   Because the cathode has a plentiful supply of electrons, the charge represented by the number of electrons released by the cathode (Q_k) is in phase with the grid voltage, as shown in Figs. 23.11(a) and 23.11(b).   The charge induced on the grid (Q_g) by these electrons would also be in phase with the grid voltage if the charges released by the cathode were to reach the plane of the grid in zero time, as shown in Fig. 23.11(c).   In this hypothetical case, the grid current due to this induced charge (Fig. 23.11(d)) leads the grid voltage by 90 degrees, because by definition, current is the time rate at which charge passes a given point.   However, the charge released by the cathode actually propagates towards the plate with finite velocity ; therefore, maximum charge is induced on the grid at a time later than that corresponding to maximum grid voltage,

as shown in Fig. 23.11(e). This condition corresponds to a shift in phase by an angle θ of Q_g with respect to e_g ; hence, the grid current lags behind the capacitive current of Fig. 23.11(d) by an angle θ, as shown in Fig. 23.11(f). Clearly, the angle θ increases with frequency and with the time of transit τ. Expressed in radians, $\theta = \omega\tau$.
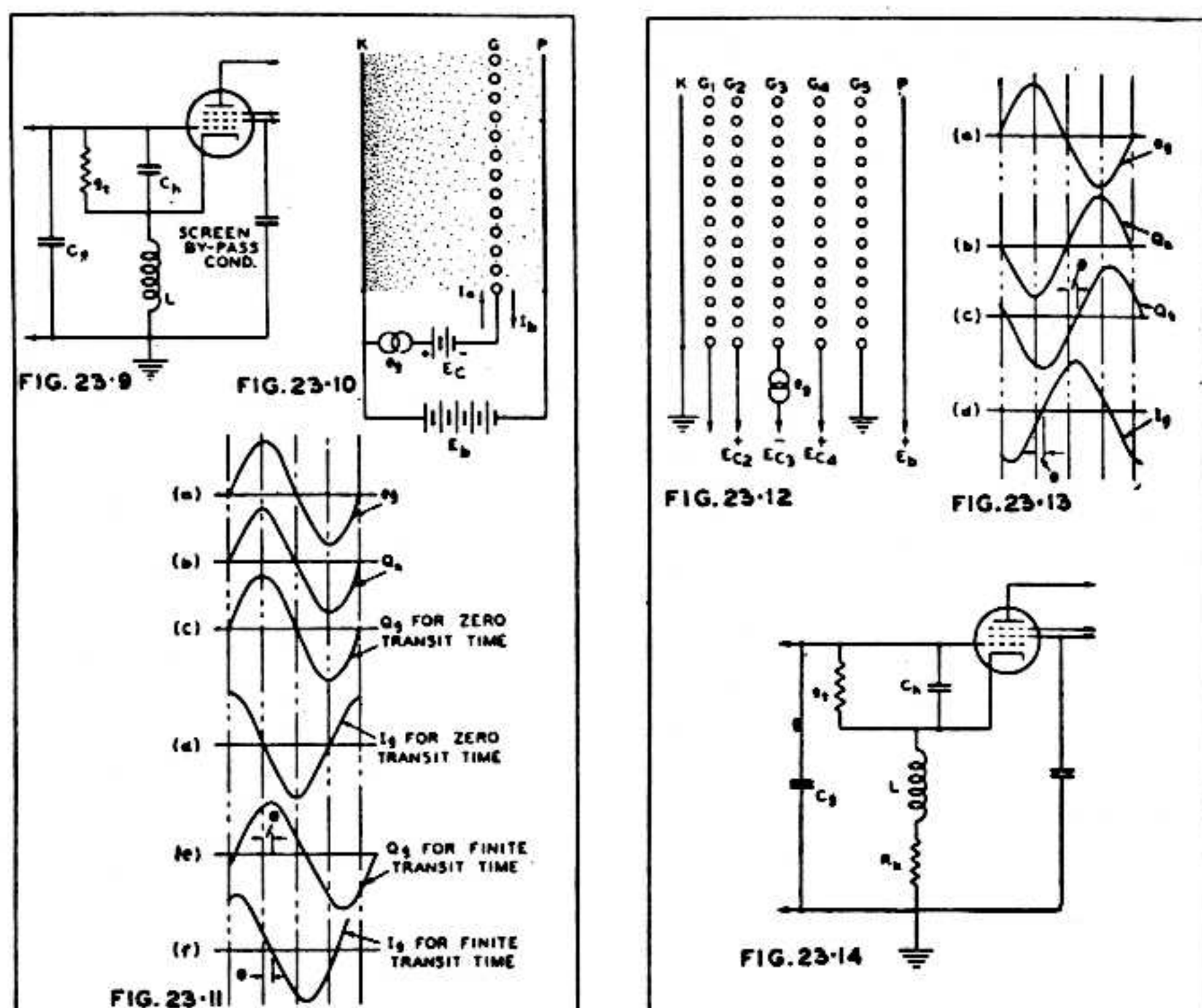
FIGURES 23·9 TO 23·14 INCLUSIVE

The amplitude of Q_g is proportional to the amplitude of the grid voltage ; the grid current, which is the time rate of change Q_g, is thus proportional to the time rate of change of grid voltage. For a sinusoidal grid voltage, $e_g = E_g \sin \omega t$, the time rate of change of grid voltage is $\omega E_g \cos \omega t$. Therefore, for a given valve type and operating point, the amplitude of grid current is

$$I_g = KE_g\omega$$

and the absolute value of grid-cathode admittance due to induced charge on the grid is

$$Y_t = I_g/E_g = K\omega \tag{3}$$

The conductive component (g_t) of this admittance is

$$g_t = Y_t \sin \theta = Y_t\theta = K\omega\theta \text{ (for small values of } \theta).$$

Because $\theta = \omega\tau$, this conductance becomes, for a given operating point,

$$g_t = K\omega^2\tau. \tag{4}$$

Thus, the conductance due to electron transit time also varies with the square of the frequency. This conductance and the input conductance, $g_m\omega^2 LC_h$, due to cathode-lead inductance, are the principal components of the term k_hf^2 of eqn. (1).

This explanation of input admittance due to induced grid charge is based on a space-charge-limited valve, and shows how a positive input admittance can result from the induced charge. The input admittance due to induced grid charge is negative in a valve which operates as a temperature-limited valve, that is, as a valve where cathode emission does not increase when the potential of other electrodes in the valve is increased. The emission of a valve operating with reduced filament voltage is temperature limited ; a valve with a screen interposed between cathode and grid acts as a temperature-limited valve when the screen potential is reasonably high. The existence of a negative input admittance in such a valve can be explained with the aid of Fig. 23.12.

When the value of E_{c2} in Fig. 23.12 is sufficiently high, the current drawn from the cathode divides between g_2 and plate ; any change in one branch of this current is accompanied by an opposite change in the other. As a first approximation, therefore, it is assumed that the current entering the space between g_2 and g_3 is constant

and equal to ρv, where ρ is the density of electrons and v is their velocity. g_2 may now be considered as the source of all electrons passing to subsequent electrodes.

Suppose now, that a small alternating voltage is connected in series with grid g_3, as shown in Fig. 23.12. During the part of the cycle when e_g is increasing, the electrons in the space between g_2 and g_3 are accelerated and their velocities are increased. Because the current ρv is a constant, the density of electrons (ρ) must decrease. In this case, therefore, the charge at g_2 is 180 degrees out of phase with the grid voltage, as shown at a and b of Fig. 23.13. This diminution in charge propagates toward the plate with finite velocity and induces a decreasing charge on the grid. Because of the finite velocity of propagation, the maximum decrease in grid charge occurs at a time later than that corresponding to the maximum positive value of e_g, as shown in Fig. 23.13(c). The current, which is the derivative of Q_g with respect to time, is shown in Fig. 23.13(d). If there were no phase displacement ($\theta = 0$), this current would correspond to a negative capacitance; the existence of a transit angle θ, therefore, corresponds to a negative conductance. By reasoning similar to that used in the derivation of eqns. 3 and 4, it can be shown that the absolute value of negative admittance due to induced grid charge is proportional to ω, and that the negative conductance is proportional to ω^2. These relations are the same as those shown in eqns. 3 and 4 for the positive admittance and positive conductance of the space-charge-limited case.

A negative value of input conductance due to transit time signifies that the input circuit is receiving energy from the " B " supply. This negative value may increase the gain and selectivity of a preceding stage. If this negative value becomes too large, it can cause oscillation. A positive value of input conductance due to transit time signifies that the signal source is supplying energy to the grid. This energy is used in accelerating electrons toward the plate and manifests itself as additional heating of the plate. A positive input conductance can decrease the gain and selectivity of a preceding stage.

It should be noted that, in this discussion of admittance due to induced grid charge, no mention has been made of input admittance due to electrons between grid and plate. The effect of these electrons is similar to that of electrons between grid and cathode. The admittance due to electrons between grid and plate, therefore, can be considered as being included in eqn. (3).

(D) Change in input capacitance

The hot grid-cathode capacitance of a valve is the sum of two components, the cold grid-cathode capacitance, C_c, which exists when no cathode current flows, and a capacitance, C_t, due to the charge induced on the grid by electrons from the cathode. The capacitance C_t can be derived from eqn. (3), where it is shown that the grid-cathode admittance due to induced grid-charge is

$$Y_t = K\omega.$$

The susceptive part of this admittance is $Y_t \cos \theta$. Since this susceptance is equal to ωC_t, the capacitance C_t is

$$C_t = K \cos \theta = K \qquad \text{(for small values of } \theta).$$

Hence, the hot grid-cathode capacitance C_h is

$$C_h = C_c + K.$$

The total input capacitance of the circuit of Fig. 23.9 when the valve is in operation, includes the capacitance C_h and a term due to inductance in the cathode lead. This total input capacitance, C_i, can be shown to be approximately

$$C_i = C_g + C_h - g_m g_t L \qquad (5)$$

where the last term shows the effect of cathode-lead inductance. This last term is usually very small. It can be seen that if this last term were made equal in magnitude to $C_g + C_h$, the total input capacitance would be made zero. However, the practical application of this fact is limited because g_m and g_t change with change in electrode voltages, and g_t changes with change in frequency.

When cathode current is zero, the total input capacitance is practically equal to $C_g + C_c$. Subtracting this cold input capacitance from the hot input capacitance given by eqn. (5), we obtain the difference, which is $K - g_m g_t L$. In general, K is

greater than $g_m g_t L$.   Therefore, in a space-charge-limited valve, where K is positive, the hot input capacitance is greater than the cold input capacitance.   In a temperature-limited valve, where K is negative, the hot input capacitance is less than the cold input capacitance.   In both valves K changes with change in transconductance. Because of this change, the input capacitance changes somewhat with change in a.v.c. bias.   In many receivers, this change in input capacitance is negligible because it is small compared with the tuning capacitances connected in the grid circuits of the high-frequency stages.   However, in high-frequency stages where the tuning capacitance is small, and the resonance peak of the tuned circuit is sharp, change in a.v.c. bias can cause appreciable detuning effect.

(E) Reduction of detuning effect

The difference between the hot and the cold input admittances of a space-charge-limited valve can be reduced by means of an unbypassed cathode resistor, R_k in Fig.
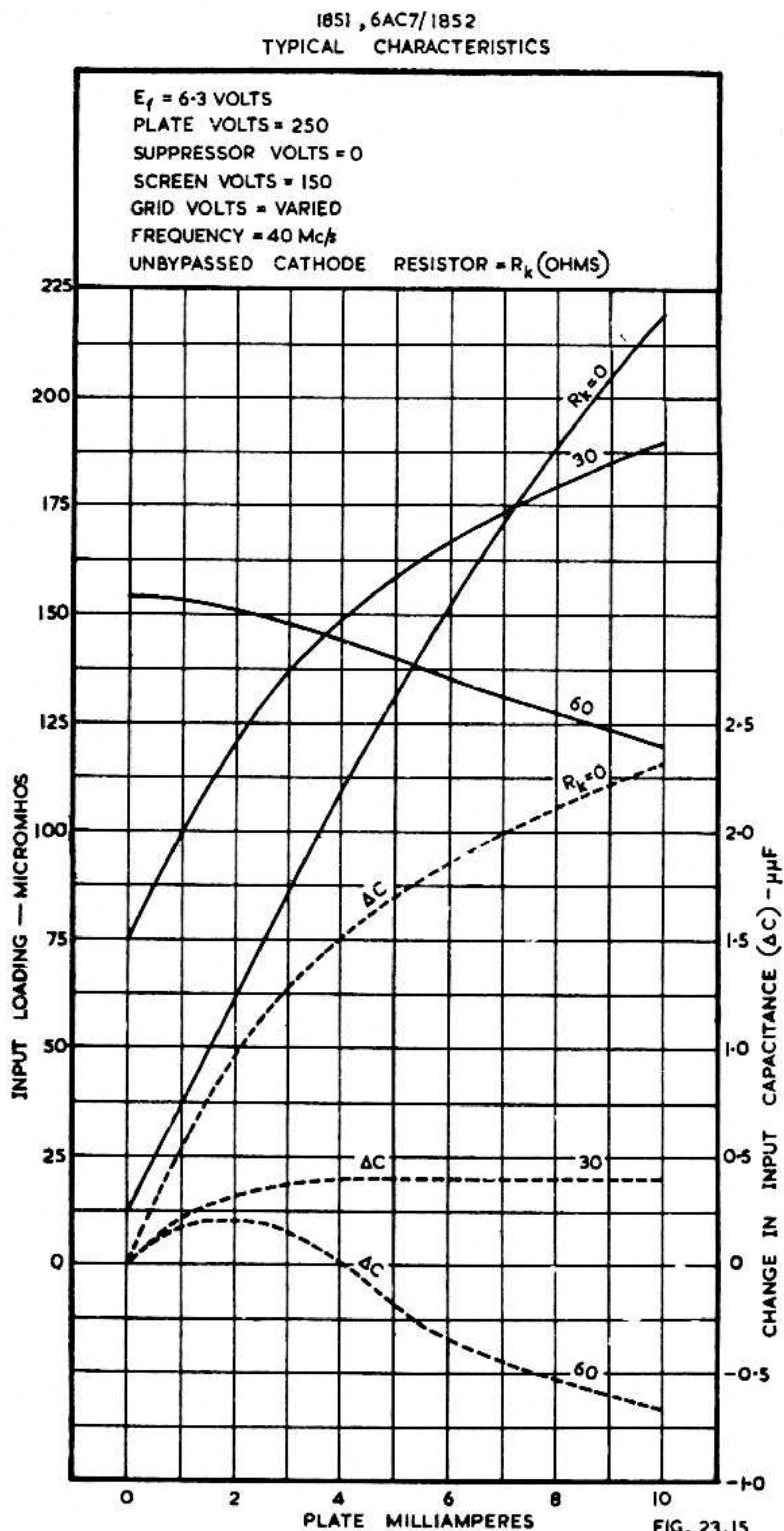
1851 , 6AC7/ 1852
TYPICAL   CHARACTERISTICS

E_f = 6·3 VOLTS
PLATE VOLTS = 250
SUPPRESSOR VOLTS = 0
SCREEN VOLTS = 150
GRID VOLTS = VARIED
FREQUENCY = 40 Mc/s
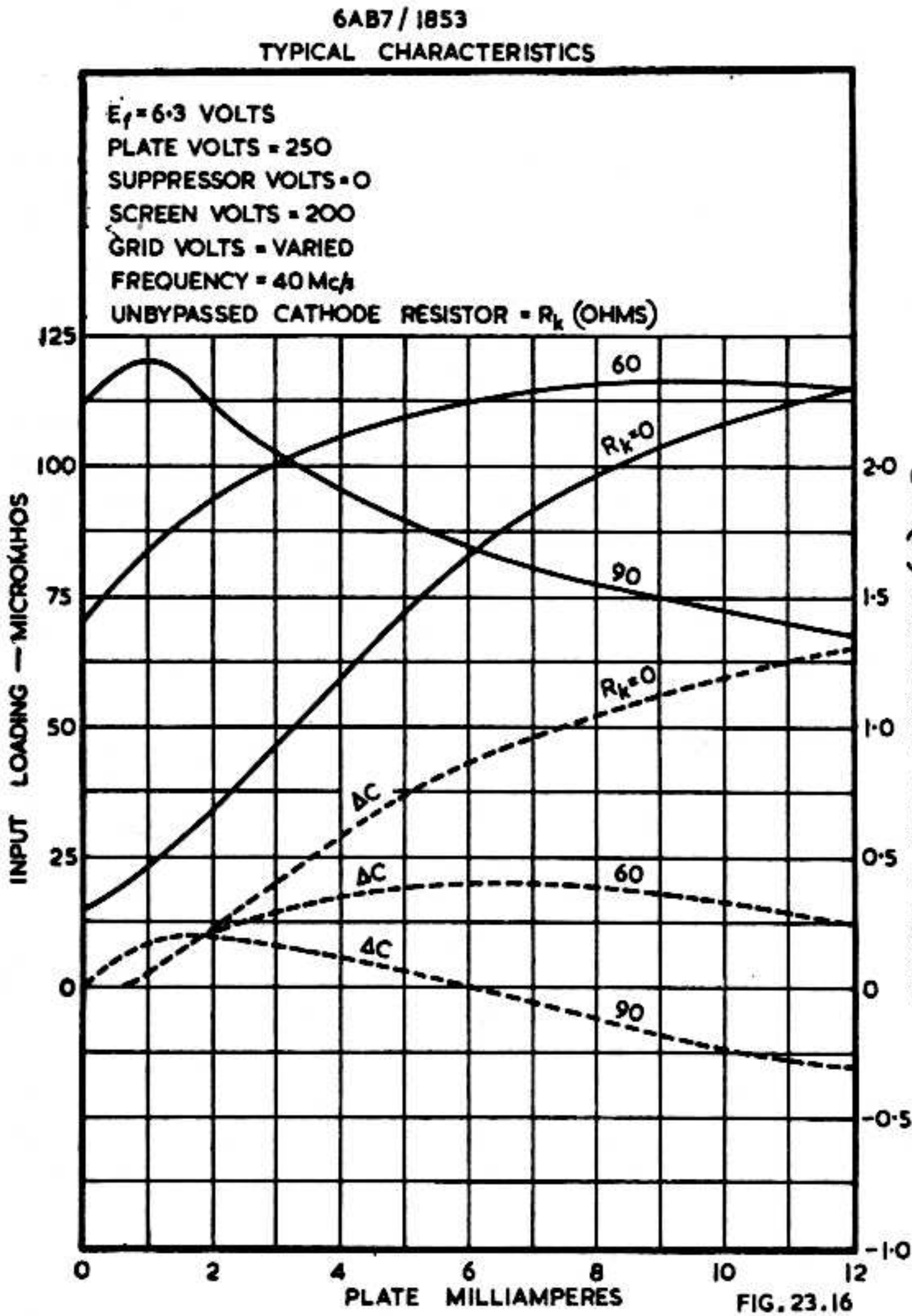UNBYPASSED CATHODE   RESISTOR = R_k (OHMS)

FIG. 23.15

23.14. The total hot input admittance of this circuit is made up of a conductance and a capacitive susceptance C_i'. Analysis of Fig. 23.14 shows that, if cathode-lead inductance is neglected, the total hot input capacitance, C_i', is approximately

$$C_i' = C_g + C_c \frac{1 + K/C_c}{1 + g_m R_k}. \tag{6}$$

Inspection of this equation shows that if K is positive and varies in proportion with g_m, the use of the proper value of R_k will make C_i independent of g_m. In a space-charge-limited valve, K is positive and is found by experiment to be approximately proportional to g_m. It follows that the proper value of R_k will minimize the detuning effect of a.v.c. in a space-charge-limited valve. Eqn. (6) is useful for illustrating the effect of R_k but is not sufficiently precise for computation of the proper value of R_k

6AB7 / 1853
TYPICAL CHARACTERISTICS

$E_f = 6.3$ VOLTS
PLATE VOLTS = 250
SUPPRESSOR VOLTS = 0
SCREEN VOLTS = 200
GRID VOLTS = VARIED
FREQUENCY = 40 Mc/s
UNBYPASSED CATHODE RESISTOR = R_k (OHMS)

FIG. 23.16

to use in practice. This value can be determined by experiment. It will be found that this value, in addition to minimizing capacitance change, also reduces the change in input conductance caused by change in a.v.c. bias. The effect of unbypassed cathode resistance on the change in input capacitance and input conductance of types 6AC7/1852 and 6AB7/1853 is shown in Figs. 23.15 and 23.16. These curves were taken at a frequency of 40 megacycles. The curves for the 6AC7/1852 also hold good for the 1851.

It should be noted that, because of degeneration in an unbypassed cathode resistor, the use of the resistor reduces gain. The reduced gain is $1/(1 + g_m R_k)$ times the gain with the same electrode voltages but with no unbypassed cathode resistance.

The hot input conductance of a valve with an unbypassed cathode resistor can be determined by modification of the values of k_h in the table. The value of k_h in the table should be multiplied by $g_m/(1 + g_m R_k)$. The resultant value of k_h, when substituted in eqn. (1), with k_c from the table, gives the input conductance of a valve with an unbypassed cathode resistor. In the factor $(1 + g_m R_k)$, g_m is the grid-cathode transconductance when R_k is by-passed.

When an unbypassed cathode resistor is used, circuit parts should be so arranged that grid-cathode and plate-cathode capacitances are as small as possible. These capacitances form a feedback path between plate and grid when there is appreciable impedance between cathode and ground. To minimize plate-cathode capacitance, the suppressor and the screen by-pass condenser should be connected to ground rather than to cathode.

SECTION 6 : VALVE AND CIRCUIT NOISE*

(i) *Thermal agitation noise* (ii) *Shot noise* (iii) *Induced grid noise* (iv) *Total noise calculations* (v) *Sample circuit calculations* (vi) *Conclusions.*

Maximum receiver sensitivity is not, in most cases, determined by the gain of the particular receiver but by the magnitude of the input circuit noise, which is generated by the antenna, the tuned input circuit, and the first tube. This is true of A-M, F-M and television except that in F-M and television the random noise effect assumes a far greater degree of importance than in the standard broadcast band. The reason for this is twofold :

(1) At the frequencies where these two services operate, 50 to 250 Mc/s, the relative values of the several different noise sources assume entirely new proportions and the heretofore unimportant and little known **induced grid noise** becomes one of the predominant components of the total.

(2) Most random input and tube noise is proportional to the square root of the bandwidth. Both television, with a 4 Mc/s band, and F-M, with a 200 Kc/s band, occupy much wider sections of the frequency spectrum than anything previously encountered by the commercial receiver engineer.

(i) Thermal agitation noise

When an alternating electric current flows through a conductor, electrons do not actually move along the conductor but they are displaced, an infinitesimal amount, first in one direction and then in the other. A voltage is built up across the conductor equal to the magnitude of the current times its resistance. Applying heat to the conducting material agitates the molecules of the conductor and, consequently, varies the instantaneous position in space of the electrons. This random electron motion is, in a sense, a minute noise current flowing through the material and is known as **thermal agitation noise.** That is, the application of heat agitates the electron distribution of the substance thereby creating the noise.

The magnitude of the short-circuit noise current is given by

$$i_n{}^2 = \frac{4KT\Delta F}{R} \tag{1}$$

where $i_n{}^2$ = mean squared noise current (amperes²)

$\quad\quad K$ = Boltzmann's Constant (Joules per degree Kelvin), 1.38×10^{-23}

$\quad\quad T$ = temperature (degrees Kelvin)

$\quad\quad \Delta F$ = bandwidth (c/s)

and $\quad R$ = resistance (ohms).

All noise currents and voltages are random fluctuations and occupy an infinite frequency band. Because of the random effect, the most convenient terminology to use in expressing their magnitude is average noise-power output. Mean-squared noise current or mean-squared noise voltage, either of which is proportional to average power, is generally used.

In the expression for various noise components the term ΔF refers to the effective bandwidth of the circuit. This is determined from a curve of power output versus frequency by dividing the area under the curve by the amplitude of the power at the noise frequency in question. For most calculations, however, where only approximate values are desired, the bandwidth between half power points, or 0.707 voltage points, will give sufficient accuracy.

The equation below expresses thermal agitation noise as a voltage in series with a given resistor ;

$$e_n{}^2 = 4KT\Delta FR. \tag{2}$$

Both the above forms are true of all resistive circuit elements or combination of elements including parallel and series-tuned circuits.

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*This section is taken directly from an article " Input Circuit Noise Calculations for FM and Television Receivers " by W. J. Stolze, published in Communications, Feb. 1947, and reprinted by special permission.

Referring to Fig. 23.17(a), let us suppose a resistance of 10 000 ohms were connected to the input of an amplifier with a 5 Kc/s bandwidth, i.e., 5 Kc/s between half power points or an audio band of 2.5 Kc/s. At room temperature, 20°C or 293°K, the term $4KT$ in the expressions for noise simplifies to 1.6×10^{-20}, which may be used in most receiver calculations. The noise in Fig. 23.17(a) is therefore :

$$e_n{}^2 = 1.6 \times 10^{-20} \Delta FR$$
$$e_n = \sqrt{1.6 \times 10^{-20} \times 5\,000 \times 10\,000}$$
$$e_n = 0.89 \text{ microvolt.}$$

The noise bandwidth is generally determined by the narrowest element in the entire circuit under consideration. In the example for Fig. 23.17(b) the bandwidth of the amplifier is narrower than the tuned circuit and therefore its ΔF is used in the calculations.
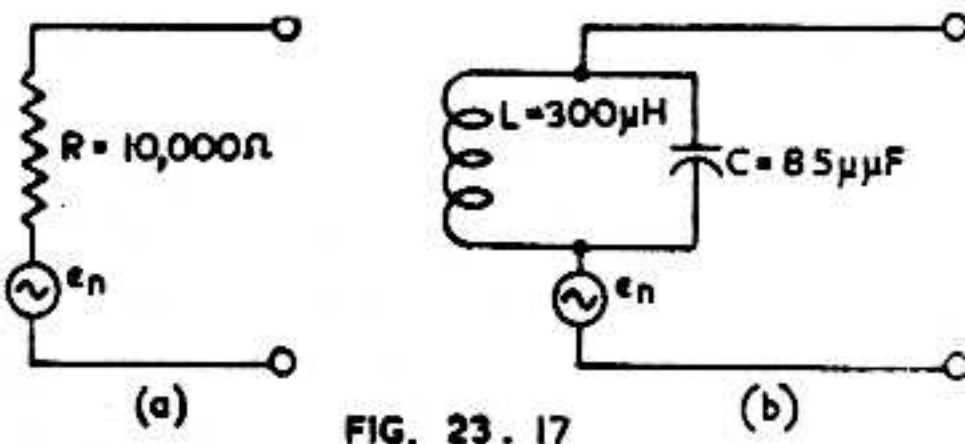

(a)   FIG. 23.17   (b)

Fig. 23.17. Sample thermal noise circuits : Frequency = 1000 Kc/s; Q = 100; L = 300 microhenries (Ref. B33).

Fig. 23.17(b) is a simple parallel-tuned circuit where the noise generating resistance is equal to the tuned circuit impedance. Again let us assume the bandwidth to be five Kc per second.

$$R = Q(\omega L) = 100 \times 1900 = 190\,000 \text{ ohms}$$
$$e_n{}^2 = 1.6 \times 10^{-20} \Delta FR$$
$$e_n = \sqrt{1.6 \times 10^{-20} \times 5\,000 \times 190\,000}$$
$$e_n = 3.9 \text{ microvolts.}$$

Thermal agitation noise voltage may be calculated easily with eqn. (2) but by using the graph shown in Fig. 23.18 the room temperature values may be found directly.
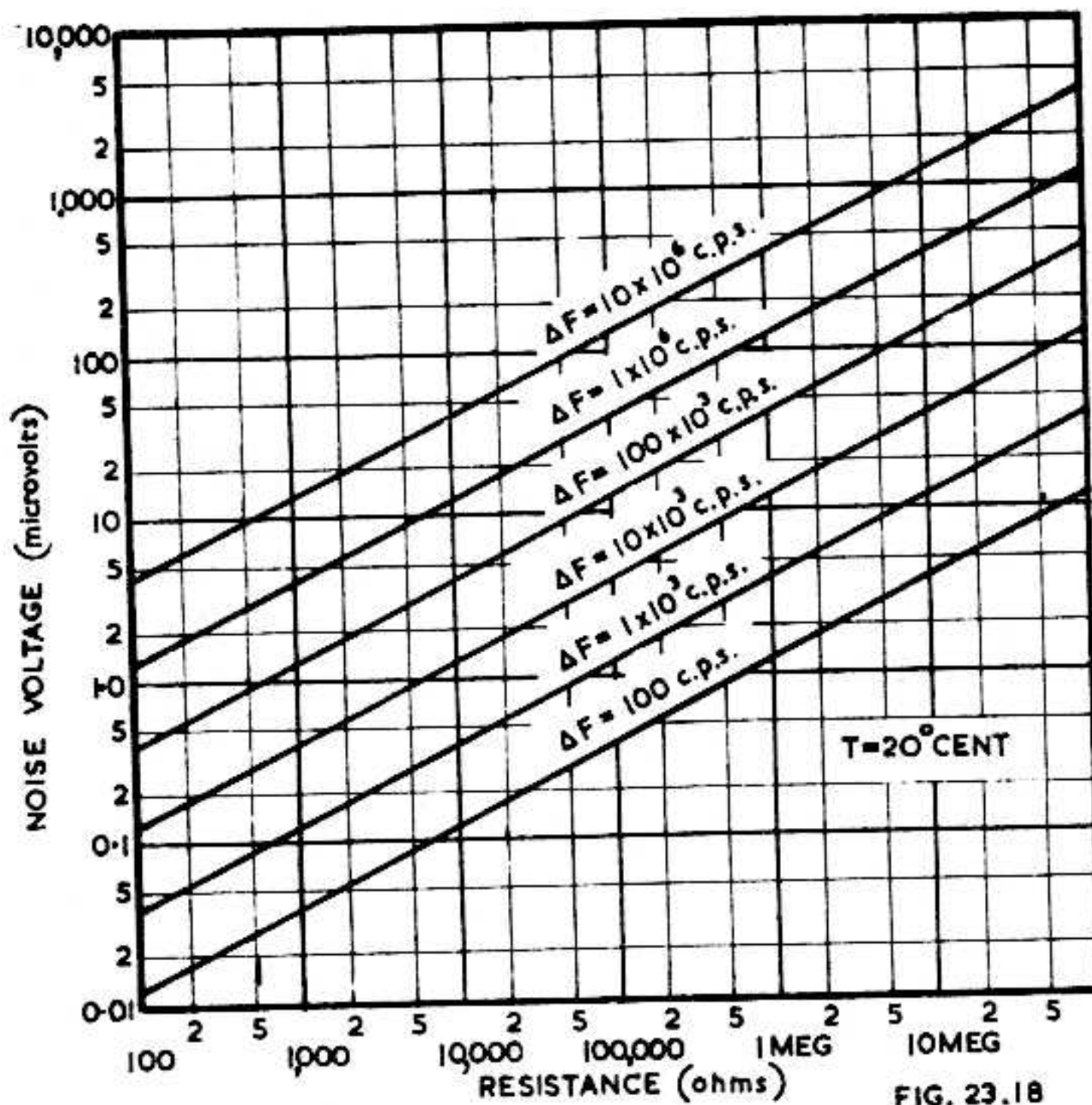

Fig. 23.18. Thermal agitation noise voltage versus resistance and band-width (Ref B33).

(ii) Shot noise

Another important component of the total receiver noise is shot noise. This noise is generated inside the vacuum tube and is due to the random fluctuations in the plate current of the tube, or, to state it in another manner, random variations in the rate of

arrival of electrons at the plate. When amplified, this noise sounds as if the plate were being bombarded with pebbles or as if a shower of shot were falling upon a metal surface, hence the name shot noise.

Although generated essentially in the plate circuit of the tube, which is not a convenient reference point for sensitivity or signal-to-noise ratio calculations, the shot noise is nearly always referred to as a noise voltage in series with the grid. Since the following equation is true,

$$e_g = i_p/g_m \tag{3}$$

where e_g = a.c. grid voltage

i_p = a.c. plate current,

and g_m = transconductance,

by simply dividing the noise current in the plate circuit by the transconductance of the tube, the shot noise may be referred to the grid and expressed in terms of grid voltage.

Another step is taken, however, to simplify the noise nomenclature. Suppose a given tube has a shot noise equal to e_n microvolts in series with its grid. It is perfectly valid to imagine that this voltage could be replaced by a resistance whose thermal agitation noise is equal to e_n (the shot noise) and to consider the tube to be free of noise. This imaginary resistance, which when placed in the grid of the tube generates a voltage equal to the shot noise of the tube, is known as the shot noise equivalent resistance or just as the equivalent noise of the tube. The advantage of this terminology is that when the equivalent noise resistance of the particular tube is known, the noise voltage may be calculated directly for any given bandwidth by substituting values in the following formula :

$$e_n^2 = 4KT\Delta F R_{eq} \tag{4}$$

where R_{eq} = equivalent noise resistance,

or at room temperature

$$e_n^2 = 1.6 \times 10^{-20} \Delta F R_{eq} \tag{5}$$

If the noise were expressed as a voltage or current its value would be correct only for one particular bandwidth.

TRIODE AMPLIFIER	$R_{EQ} = \dfrac{2 \cdot 5}{G_M}$
PENTODE AMPLIFIER	$R_{EQ} = \dfrac{I_B}{I_B + I_{G_2}} \left(\dfrac{2 \cdot 5}{G_M} + \dfrac{20 I_{G_2}}{G_M^2} \right)$
TRIODE MIXER	$R_{EQ} = \dfrac{4}{G_C} \qquad G_C = \dfrac{G_M}{4}$
PENTODE MIXER	$R_{EQ} = \dfrac{I_B}{I_B + I_{G_2}} \left(\dfrac{4}{G_C} + \dfrac{20 I_{G_2}}{G_C^2} \right)$
MULTIGRID CONVERTER OR MIXER	$R_{EQ} = 20 \dfrac{I_B (I_K - I_B)}{I_K G_C^2}$

R_{EQ} = EQUIVALENT SHOT NOISE RESISTANCE I_{G_2} = AVERAGE SCREEN CURRENT

G_M = GRID PLATE TRANSCONDUCTANCE G_C = CONVERSION TRANSCONDUCTANCE

I_B = AVERAGE PLATE CURRENT I_K = AVERAGE CATHODE CURRENT

FIG. 23.19

Fig. 23.19. Approximate calculated equivalent noise resistance of various receiving-type tubes (Ref. B33).

By knowing the R_{eq} of any two given tubes their relative shot noise merit is also known regardless of what bandwidth they are to operate at, while if the noise voltages were given alone the operating bandwidth at which the calculation was made would also have to be noted if the relative merits of the two tubes were to be defined.

Noise-equivalent resistance values for a number of different tube types (triodes, pentodes, and converters) and for various circuit applications (amplifiers and mixers) can be calculated by applying the expressions presented in the chart, Fig. 23.19*.

When the term converter is used it refers to a tube that is used for frequency conversion where the single tube acts as the local oscillator and the mixer (6SA7) ; the term mixer where two tubes are used, one as the mixer (6SG7), and one as the local oscillator (6C4).

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*W. A. Harris " Fluctuations in vacuum tube amplifiers and input systems," R.C.A. Review, April 1941.

TUBE TYPE	APPLICATION	PLATE VOLTS*	SCREEN VOLTS	TRANSCONDUCTANCE MICROMHOS	EQUIVALENT NOISE RESISTANCE OHMS
6AC7	PENTODE AMPLIFIER	300	150	9,000	720
6AC7	PENTODE MIXER	300	150	2,200	2,800
6AG5	PENTODE AMPLIFIER	250	150	5,000	1,650
6AG5	PENTODE MIXER	250	150	1,250	6,600
6AG7	PENTODE AMPLIFIER	300	150	11,000	1,540
6AK5	PENTODE AMPLIFIER	180	120	5,100	1,880
6AK5	PENTODE MIXER	180	120	1,280	7,520
6AK6	PENTODE AMPLIFIER	180	180	2,300	8,800
6AT6	TRIODE AMPLIFIER	250	—	1,200	2,100
6AU6	PENTODE AMPLIFIER	250	150	5,200	2,660
6BA6	PENTODE AMPLIFIER	250	100	4,400	3,520
6BA6	PENTODE MIXER	250	100	1,100	14,080
6BE6	CONVERTER	250	100	475 •	190,000
6C4	TRIODE AMPLIFIER	100	—	3,100	810
6C4	TRIODE MIXER	100	—	770	3,240
6C5	TRIODE AMPLIFIER	250	—	2,000	1,250
6C5	TRIODE MIXER	250	—	500	5,000
6J5	TRIODE AMPLIFIER	250	—	2,600	960
6J5	TRIODE MIXER	250	—	650	3,840
½6J6	TRIODE AMPLIFIER	100	—	5,300	470
½6J6	TRIODE MIXER	100	—	1,320	1,880
6K8	CONVERTER	250	100	350 •	290,000
6SA7	CONVERTER	250	100	450 •	240,000
6SB7-Y	CONVERTER	250	100	950 •	62,000
6SC7	TRIODE AMPLIFIER	250	—	1,325	1,890
6SG7	PENTODE AMPLIFIER	250	125	4,700	3,100
6SG7	PENTODE MIXER	250	125	1,180	12,400
6SJ7	PENTODE AMPLIFIER	250	100	1,650	6,100
6SK7	PENTODE AMPLIFIER	250	100	2,000	11,000
6SL7	TRIODE AMPLIFIER	250	—	1,600	1,560
6SQ7	TRIODE AMPLIFIER	250	—	1,100	2,300

(*) VALUES OF PLATE VOLTAGE AND CURRENT AND SCREEN VOLTAGE AND CURRENT ARE FOR TYPICAL OPERATING CONDITIONS.

(•) CONVERSION TRANSCONDUCTANCE - MICROMHOS

FIG. 23.20

After the equivalent noise resistance is known the value of r.m.s. noise voltage at the grid of this tube can be calculated by applying the same expression that is used for thermal agitation noise,

$$e_n{}^2 = 1.6 \times 10^{-20} \Delta FR$$

or, by using the graph of Fig. 23.18.

Fig. 23.20 presents calculated equivalent noise resistance values for a number of commonly used tubes acting as various types of circuit elements. These are, of course, approximate figures.

It can be seen from Figs. 23.19 and 23.20 that the noise resistance or voltage is at a minimum for a triode, increasing for the pentode and the multigrid tube, following in that order.

Shot noise is unique among the noise sources in the sense that the shot-noise voltage should be considered to exist in series with the grid inside the tube. The reason for this is that nothing can be done to the external grid circuit that will alter the magnitude of this component. Even though the shot noise must be tolerated, its effect can be minimized by designing the input circuit for maximum signal at the grid. This does not reduce the magnitude of the noise but does improve the signal-to-noise-ratio of the receiver.

(iii) Induced grid noise

Also present in the receiving tube is a third source of noise which is generated internally in the tube but whose magnitude and effect are determined partially by the external input circuit. Known as induced grid noise, this minute current is induced in the grid wires of the tube by random fluctuations in the plate current. It is known that a varying electron beam will induce a current in any nearby conductor. Therefore, the fluctuating plate current which is in a sense a varying electron beam, will induce a noise current in the nearby grid conductors.

The input impedance of a vacuum tube has a reactive and a resistive component. At relatively low frequencies the resistive component is very high (below about 30 Mc/s) ; as the frequency is increased the resistive component decreases and its magnitude eventually becomes comparable to or even lower than the external grid circuit impedance. The resistive component is composed of two parts, the portion due to transit time effect, and the portion due to the inductance of the cathode lead.
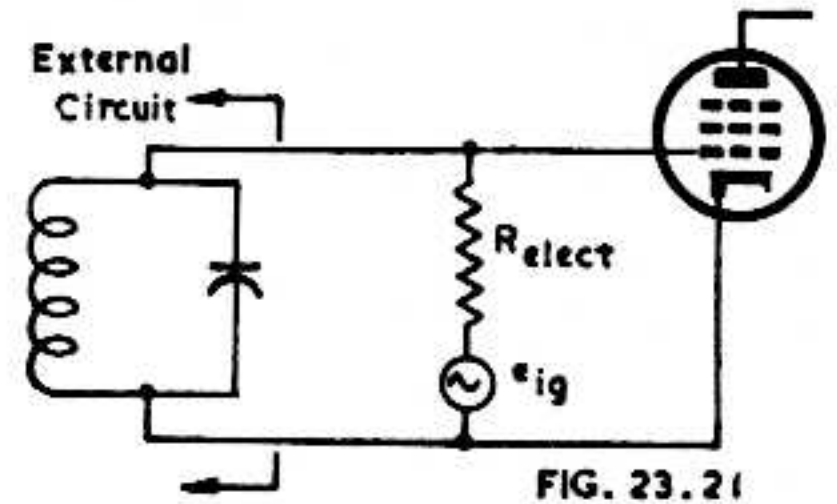
An expression for induced-grid-noise* for tubes

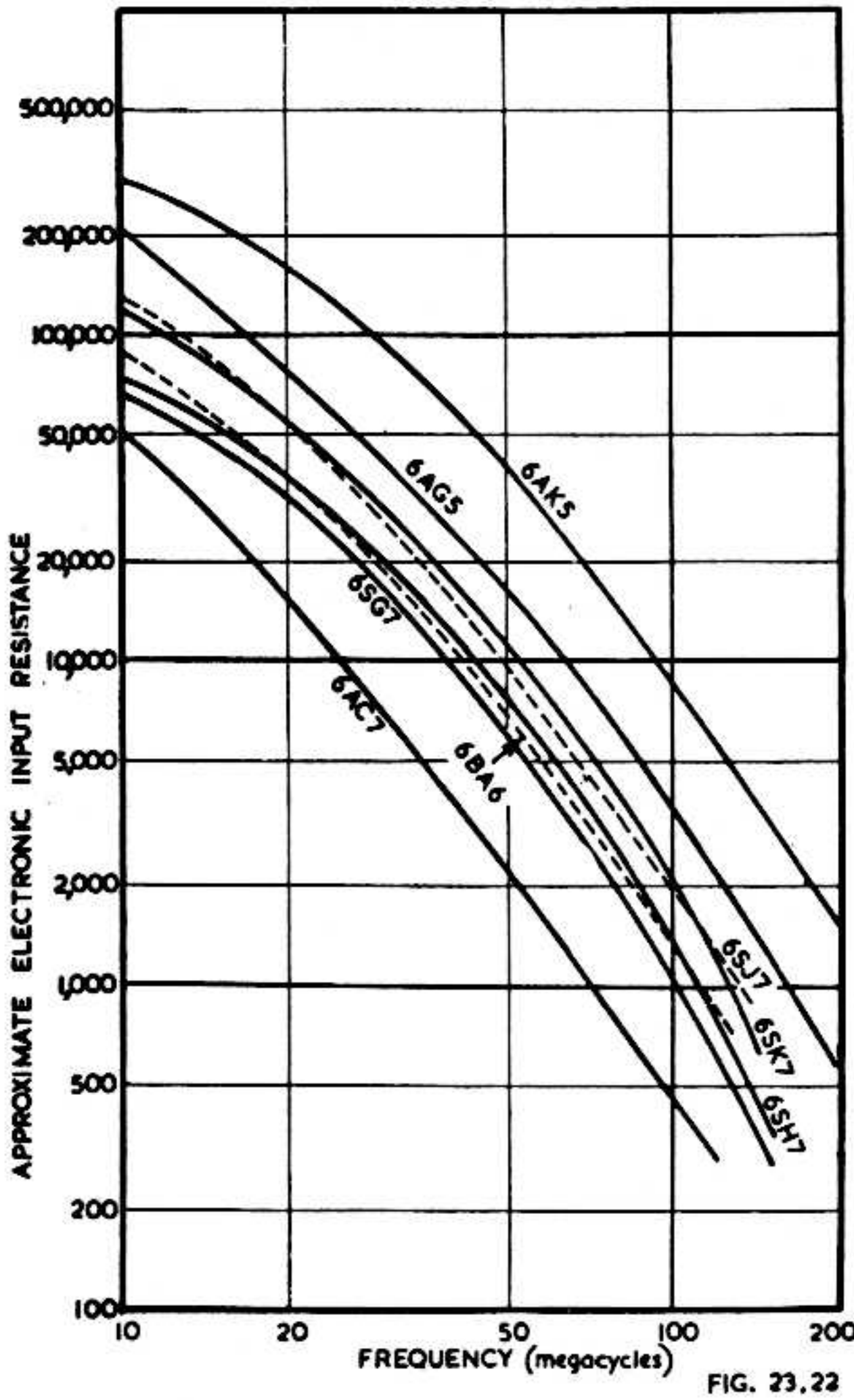Fig. 23.21. *Position of induced grid noise in vacuum-tube circuit (Ref. B33).*

Fig. 23.22. *Approximate electronic input resistance versus frequency (Ref. B33).*

with control grid adjacent to the cathode follows :

$$i^2_{i.g.} = 1.4 \times 4KT_k \Delta F G_{elect}$$

or when expressed in the form of a voltage generator,

$$e^2_{i.g.} = 1.4 \times 4KT_k \Delta F R_{elect} \quad (6)$$

where : T_k = cathode temperature (degrees Kelvin)

G_{elect} = electronic (transit time) component of input conductance

and R_{elect} = electronic component of input resistance.

From eqn. (6) it can be seen that the induced grid noise is proportional to the electronic or transit time component of the input resistance. Measurement of the total input resistance is a comparatively simple matter with the use of a high frequency Q meter, but the separation of the electronic and the cathode inductance components, which are essentially two resistances in parallel between the grid and ground, is a very difficult matter. Since most high-frequency tubes are constructed with either two cathode leads or one very short lead, assuming the total measured input resistance to be electronic would not introduce too great an error. Another factor in favour of this approximation is that it would be the case for maximum induced grid noise and any error introduced would more than likely be on the safe side.

*D. O. North " Fluctuations induced in vacuum-tube grids at high frequencies " Proc. I.R.E. Feb. 1941.

Cathode temperature in most receiving tubes, which almost exclusively use oxide-coated cathodes, is approximately 3.6 times the normal room temperature in degrees K. Eqn. (6) can be rewritten therefore as

$$e^2{}_{i.g.} = 5 \times 4KT\Delta FR_{elect} \tag{7}$$

where T = room temperature (degrees Kelvin),
or, when T = 300 degrees Kelvin,

$$e^2{}_{i.g.} = 8 \times 10^{-20}\Delta FR_{elect} \tag{8}$$

In circuit calculations this noise is essentially in series with a resistance equal to R_{elect} located between the grid and ground—Fig. 23.21.

The approximate input resistance for a number of common receiving tubes in the frequency range of F-M and television is given in Fig. 23.22. This chart can be used to find approximate input resistance values for induced grid-noise calculations.
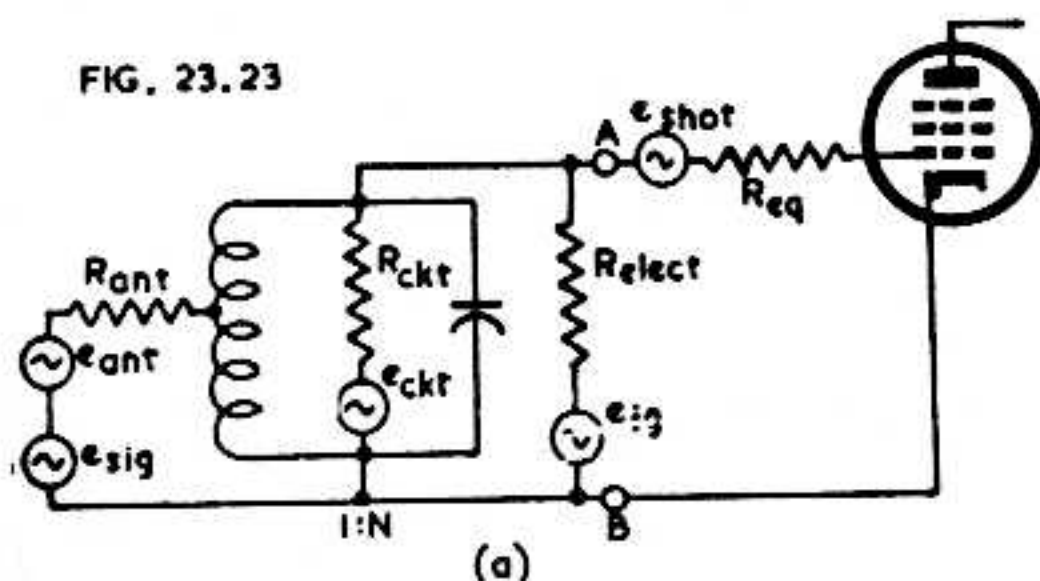
(iv) Total noise calculations

Calculations of total input noise are made by using the grid of the input tube as a reference point. There are many sources of noise and each must be calculated and referred to the grid reference point before a summation is made. Since noise is a random effect and calculated on a power basis, the separate components cannot be added directly but as the square root of the sum of the squares.

$$\text{Total Noise} = \sqrt{e_1{}^2 + e_2{}^2 + e_3{}^2 + \text{etc.}} \tag{9}$$

The various noise voltages that must be referred to the first grid are :
(1) Thermal agitation noise of the antenna radiation resistance.
(2) Thermal agitation noise of the tuned grid circuit.
(3) Shot noise of the input tube.
(4) Induced grid noise of the input tube.
(5) Grid circuit noise of the following stages referred back to the first grid.

In Fig. 23.23(a) appears a diagram of a practical input circuit and the location of all the circuit parameters and noise voltages. Fig. 23.23(b) is essentially the same except that the antenna circuit is reflected through the transformer and considered to exist at the grid. This is the diagram that is most useful in calculating the total input circuit noise.

The steps necessary to find specific values for each of these factors are shown in Fig. 23.24. Antenna radiation resistance varies widely with the type of antenna chosen, but for F-M and television work it is generally in the order of 75 to 300 ohms. When the noise is known in terms of an equivalent resistance, as is the case here for the antenna, tuned circuit, and shot noise, the equivalent voltage can be either calculated or obtained directly from Fig. 23.18.

In order to add the antenna, tuned circuit, and induced grid noise to the shot noise the effective voltage of these three components at the grid, or between the points A and B, must be known. Each must go through what is essentially a resistive divider and may be calculated as shown in Fig. 23.25.

After knowing the magnitude of the separate sources that exist between A-B, the total noise voltage is

FIG. 23.23

R_{ant} = Antenna Radiation Resistance
e_{ant} = Antenna Resistance Thermal Noise
e_{sig} = Signal Voltage
R_{ckt} = Tuned Circuit Impedance
e_{ckt} = Tuned Circuit Thermal Noise
R_{elect} = Electronic Input Resistance
e_{ig} = Induced Grid Noise
R_{eq} = Equivalent Shot Noise Resistance
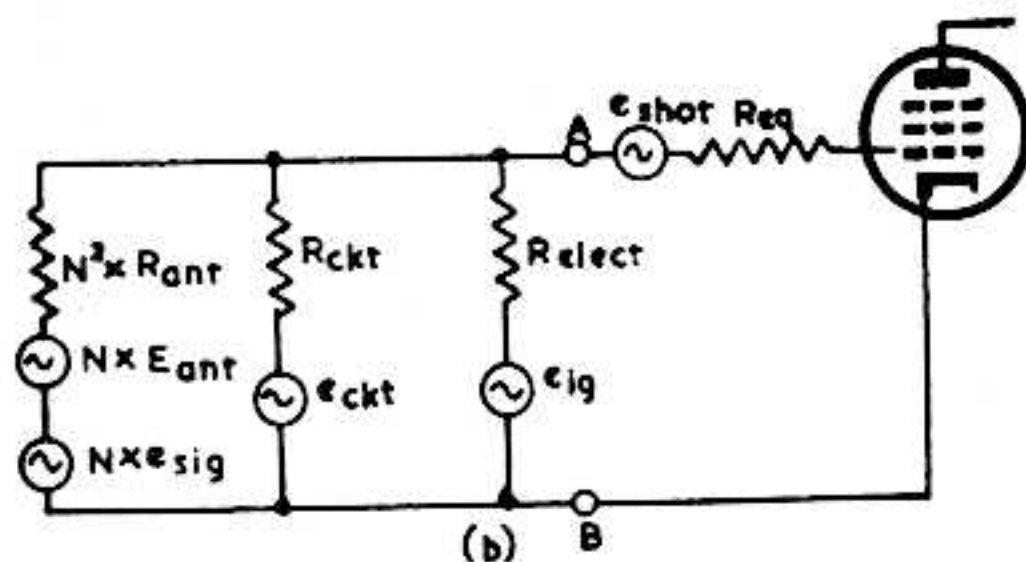e_{shot} = Valve Shot Noise
N = Turns Ratio of Coupling Transformer

Fig. 23.23. Position of various noise sources in input circuit (Ref. B33).

(1) R_{ANT}-DEPENDS UPON SPECIFIC ANTENNA

(2) $E_{ANT} = \sqrt{1.6 \times 10^{-20} R_{ANT} \Delta F}$ — OR DIRECTLY FROM FIG. 23-18

(3) $R_{CKT} = Q \omega L = \dfrac{Q}{\omega C}$

(4) $E_{CKT} = \sqrt{1.6 \times 10^{-20} R_{CKT} \Delta F}$ — OR DIRECTLY FROM FIG. 23-18

(5) R_{ELECT}-FROM ACCOMPANYING CHART, FIG. 23-22

(6) $E_{IG} = \sqrt{8 \times 10^{-20} R_{ELECT} \Delta F}$

(7) R_{EQ} -FROM ACCOMPANYING CHART, FIG. 23-20

(8) $E_{SHOT} = \sqrt{1.6 \times 10^{-20} \Delta F \, R_{EQ}}$ — OR DIRECTLY FROM FIG. 23-18

FIG. 23.24

Fig. 23.24. *Procedure for calculating various noise voltages (Ref. B33).*

$$e_{total} = \sqrt{(e_{shot})^2 + (e_{ant} \text{ at A} - \text{B})^2 + (e_{i-g} \text{ at A-B})^2 + (e_{ckt} \text{ at A-B})^2} \qquad (10)$$

One other factor may affect this total, however.   If the total noise of the following stages, which is calculated similarly, ignoring the antenna of course, is appreciable, it must be added to the constants of Fig. 23.25.   In reflecting it to the first grid the second stage noise should be divided by the gain of the first tube.   When the gain is about ten or more this factor may usually be neglected.

Effective signal voltage across A-B is calculated in the same way as the antenna noise in Fig. 23.25.   The signal-to-noise ratio is now also known.

Since the signal-to-noise ratio is determined by the signal strength and the total noise at the grid of the input tube, for a receiver that has a mixer, such as 6SK7, for the input tube, the signal-to-noise ratio may be considerably improved by the addition of an r-f tube, such as a 6SG7, which has considerably less total noise.   By adding additional r-f tubes (6SG7's), however, since the total noise and signal at the grid will be the same, the signal-to-noise ratio will not be improved.
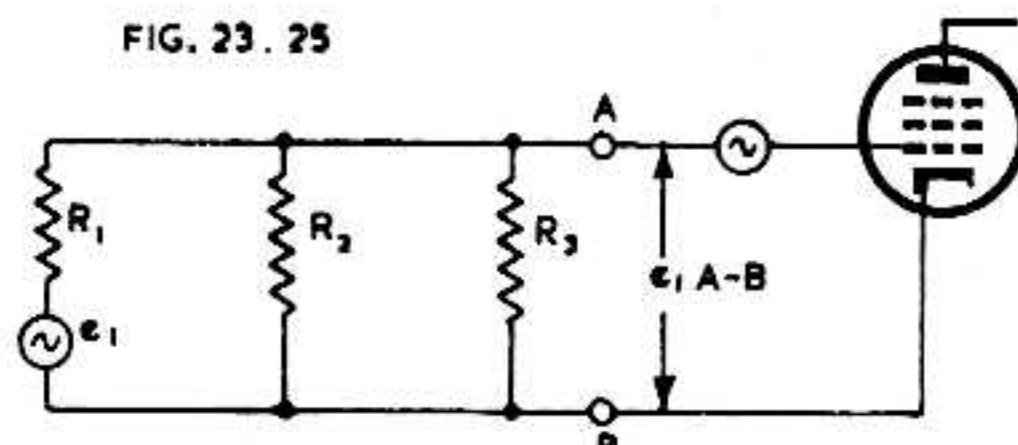
FIG. 23 . 25

Fig. 23.25.   *Circuit for reflecting various voltages to the grid (Ref. B33).   To find the effective voltage of the antenna, the tuned circuit, and the induced grid noise at the grid of the tube let R_1 equal one of the above noise resistances and e_1 its generated voltage.   If R_2 and R_3 equal the other two noise resistances the effective voltage at the grid is*

$$e_{1A-B} = \frac{e_1}{R_1 + \dfrac{R_2 R_3}{R_2 + R_3}} \times \frac{R_2 R_3}{R_2 + R_3}$$

This calculation must be performed for the three components in turn.

(v) Sample circuit calculations

For a sample problem let us calculate the total noise at the grid of an F-M receiver r-f amplifier stage, assuming the circuit in Fig. 23.26(a) to be under consideration.

As a simplification of procedure the steps in the calculations will be numbered.

(1) $N^2 R_{ant}$ = 1200 ohms (calculated)

(2) R_{elect} = 1200 ohms (Figure 23.22)

(3) R_{ckt} = $Q \omega L$ = 8000 ohms (calculated)

(4) R_{eq} = 3100 ohms (Figure 23.20)

At this point it will be convenient to redraw the circuit as shown in Fig. 23.26(b).

(5) Ne_{ant} = 2 microvolts (Fig. 23.18)

(6) $e_{i.g.}$ = $\sqrt{8 \times 10^{-20} \times 200 \times 10^3 \times 1200}$

= 4.4 microvolts (equation (8))

(7) e_{ckt} = 6 microvolts (Fig. 23.18).
(8) e_{shot} = 3.5 microvolts (Fig. 23.18).

The next step is to find the effective voltage of each source between the grid and ground (or A-B) as shown in Fig. 23.25.

$$(9)\ e_{ant}\ \text{A-B} = \frac{2}{1200 + 1040} \times 1040 = 0.93 \text{ microvolt.}$$

$$(10)\ e_{i.g.}\ \text{A-B} = \frac{4.4}{1200 + 1040} \times 1040 = 2.0 \text{ microvolts.}$$

$$(11)\ e_{ckt}\ \text{A-B} = \frac{6}{8000 + 600} \times 600 = 0.42 \text{ microvolts.}$$

and the total noise is therefore

$$(12)\ e_{total} = \sqrt{3.5^2 + 0.93^2 + 2.0^2 + 0.42^2}$$
$$= 4.3 \text{ microvolts [equation (10)].}$$

(vi) Conclusions

Selection of an input tube for a television or F-M receiver is dependent upon many varying circuit conditions and individual requirements. The choice of using balanced or unbalanced input, permeability or capacitor tuning, noisy pentodes or quiet triodes that possibly require neutralization, among others, lies entirely with the design engineer. Considering these reasons and various engineering and economic compromises no particular tube can be chosen and defined as **the input tube.** Complete noise information about the circuits involved is necessary, however, as this is one of the determining factors for good sensitivity and signal-to-nois ratio
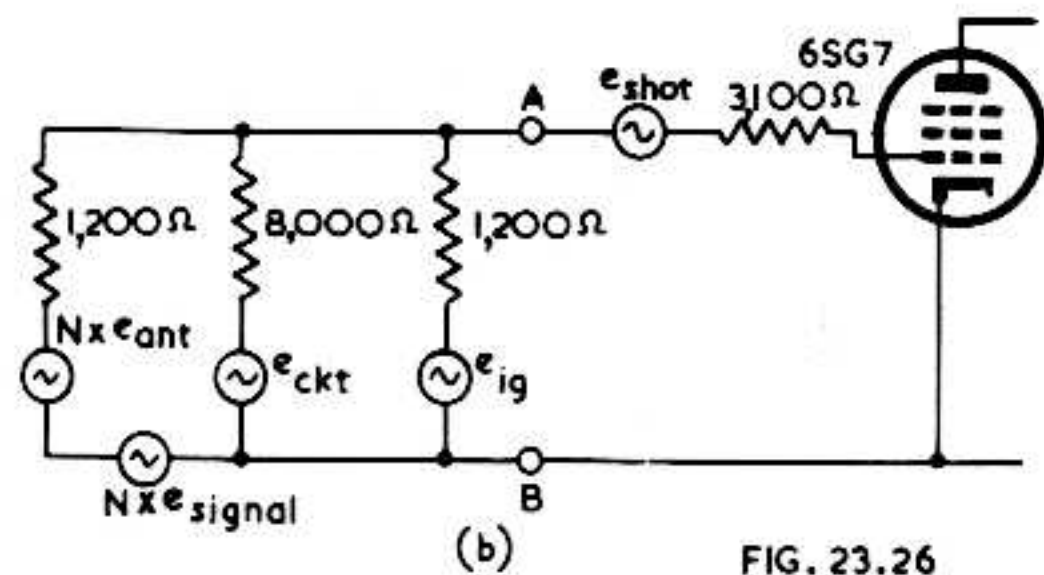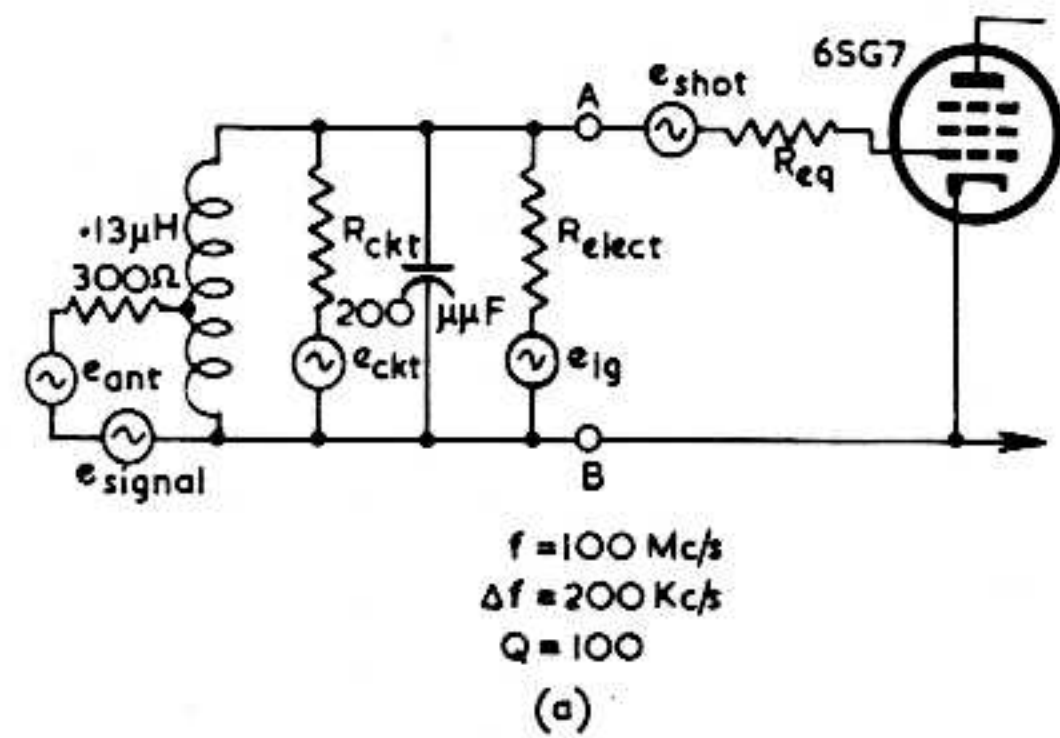

Fig. 23.26. *Typical F-M receiver input circuit (Ref. B33).*

SECTION 7 : INSTABILITY IN R-F AMPLIFIERS

(i) Causes of instability   (ii) Inter-electrode capacitance coupling   (iii) Summary.

(i) Causes of instability

Instability in r-f amplifiers can be due to many causes and some of these are listed below.

(1) Inter-electrode coupling due to capacitances within the r-f amplifier valve. This coupling may be augmented by additional capacitance external to the valve, due to wiring etc.

(2) Coupling between r-f and aerial coils, and leads, due to lack of adequate shielding or care in placement of the coils and leads relative to one another.

(3) Impedances common to several stages such as the metal shaft of a variable capacitor, power supply impedance (including heater leads), capacitances between switch contacts, a.v.c. line etc.

(4) Feedback at the intermediate frequency or on harmonics of the i-f.

Feedback at i-f can be serious when there is only an aerial stage preceding the converter. In some cases a simple expedient to overcome this difficulty is to connect a series resonant i-f trap between the aerial and earth terminals.

(5) Overall feedback from the a-f section into the aerial stage. This can be checked by bringing the speaker leads into close proximity with the aerial terminal, and the feedback usually manifests itself as a characteristic a-f howl. Practically all of the above factors are under the receiver designer's control even though the elimination of undesired oscillations is often a very difficult practical problem. Further discussion is given in Chapter 35, Sect. 3(v).

(ii) Inter-electrode capacitance coupling

Coupling due to inter-electrode capacitances calls for special consideration since these are an irreducible minimum when due care has been taken with the external circuits. A description of the effects which may be expected has been given in some detail in Chapter 2 Sect. 8 (see also Chapter 26, Sects. 7 and 8), where the nature of the impedance reflected into the grid circuit has been discussed. The connection with circuit instability is largely bound up in the magnitude of the input resistance component of the reflected impedance appearing across the valve grid circuit, and as to whether this input resistance is positive or negative. If the resistance is positive the dynamic impedance of the grid input circuit is lowered and there is a loss in gain and a broadening of the tuning characteristic. For a negative input resistance component the opposite effects hold with the dynamic impedance of the tuned circuit increased and the tuning becoming sharper. When the negative resistance, due to coupling by inter-electrode capacitances from other circuits associated with the valve, equals or is less than the positive resistance of the grid input circuit, oscillation will occur.

In the usual case for a r-f amplifier the primary winding of the r-f coil is resonated outside the tuning range. For a high impedance primary resonated below the lowest tuning frequency the plate circuit of the valve acts as a capacitance and the grid circuit is affected as though a positive resistance and a capacitance were connected in parallel across it. This effect is due to the total grid-to-plate capacitance. When the primary of the r-f coil is resonated above the highest tuning frequency, the valve sees an inductive load and a capacitance and a negative resistance component appear in parallel across the grid input circuit. (This circuit is the aerial coil when only one r-f stage is used).

For the case of the tuned anode load circuit i.e. a single tuned circuit connected directly between plate and B+, the valve sees a resistive load and there is only a capacitance effectively reflected in parallel with the valve grid input circuit. In this latter case, however, because of circuit mistracking, the valve may only see a resistive load at a few points in the tuning range and at other settings of the tuning dial the load may appear as either a capacitance or an inductance depending on whether the resonant frequency is below or above the required signal frequency.

The above effects can be neutralized by suitable circuit arrangements but usually the additional trouble and expense involved are avoided whenever possible.

To obtain a quantitive idea of the permissible values of grid-to-plate capacitance which would just put a circuit on the verge of instability, an investigation was made by Thompson (Ref. B41) who gave the following results.

$$C_{gp} = \frac{A}{\omega g_m R_D{}^2}$$

where C_{gp} = total capacitance grid to plate (valve internal and external)

$$A = \begin{cases} 2 & \text{for 1 r-f stage} \\ 1 & \text{for 2 r-f stages} \\ 0.764 & \text{for 3 r-f stages} \end{cases}$$

$\omega = 2\pi \times$ operating frequency

g_m = mutual conductance of r-f amplifier valve (assumed same type and operating conditions in each case),

and R_D = dynamic impedance of the input and output circuits (assumed identical in all cases for simplicity).

Obviously, even for values of C_{gp} less than those obtained from the above expression, there would be a marked effect on the gain and selectivity of the circuits before oscillation actually started.

(iii) Summary

Possible sources of feedback giving rise to instability can often be predicted from the circuit and component layout diagrams of a receiver.   An estimate can be made of the possible magnitudes of many of the undesired voltages involved.

It should be evident that it is far better to avoid possible feedback and instability by good electrical and mechanical design rather than spend many fruitless hours tracking down an oscillation which could have been avoided.

A most helpful discussion of instability problems is given in E. E. Zepler's " Technique of Radio Design " (in particular Chapter 9) and the reader is recommended to consult this book as an excellent practical guide.

Some further considerations will be given to circuit instability in Chapter 26, in connection with i-f amplifiers.   These circuits, being fixed tuned, are usually more amenable to calculation of possible instability than r-f stages and the results obtained more closely approximate to the practical set-up.

It is necessary to mention that at very high frequencies (say above 50 Mc/s or so) the inductance of the screen-grid lead in screen-grid tetrodes and pentodes can cause an apparent change in plate to control-grid capacitance which is often sufficient to cause instability.   If the screen is earthed by means of a capacitor, it is often possible to select a capacitance value which will be series resonant with the screen lead inductance, at the working frequency, and so form a low impedance path to ground. This arrangement is often sufficient to prevent instability from this cause, even over a range of frequencies.   For further discussion see Ref. A8.

It is also important to note that at the higher frequencies a capacitor does not behave as a pure capacitance, and its effective inductance and resistance become increasingly important as the frequency increases.   With many types of capacitors it is possible that the inductive reactance will exceed the capacitive reactance even at frequencies as low as 30 Mc/s (with electrolytics an additional r-f by-pass should always be used). This effect is sometimes used to make the capacitor series resonant at the working frequency.   Even if the capacitor behaves as a small inductance it will be appreciated that in some cases the reactance can be very low and effective by-passing is still possible.   In tuned circuits the Q can be materially affected by the increase in r-f resistance of the capacitors and both this and the previous effects should be checked.

The effectiveness of by-pass capacitors and their Q at the working frequency can be readily checked with the aid of a Q meter by placing the capacitor in series with a coil of known inductance and Q.   See Ref. B45.   Finally, r-f chokes do not always behave as such, and their resonant frequencies should always be checked (see also Chapter 11 Sect. 6).

SECTION 8 : DISTORTION

(i) Modulation envelope distortion   (ii) Cross modulation distortion.

Modulation envelope distortion introduced by r-f voltage amplifiers is usually small when compared with that introduced by the later stages in a radio receiver e.g. the i-f amplifier, the detector and, perhaps, the frequency converter.   This is particularly true when the magnitude of the signal voltage is small.   Amplitude modulation distortion in r-f amplifiers is usually caused by the curvature of the valve characteristic relating control grid voltage to plate current.   The usual explanation for the introduction of this distortion is to consider that the portion of the characteristic curve being used can be represented by a power series relating the grid voltage (in this case the modulated carrier wave) and the plate current.   It can be shown from this treatment that second order terms (terms containing squares) in the series do not introduce

distortion because of the selectivity of the tuned circuits. Third order terms (those containing cubes) do introduce distortion of the modulation envelope.

R-F amplifier valves are usually designed so that the third order curvature of their characteristics is minimized as far as possible. This is achieved by using a variable pitch for the control grid winding and results in the well known variable-mu (or remote cut-off) characteristic. The variable-mu (or more exactly " variable g_m " characteristic) and the resultant shape of the $e_g - i_p$ curve, have a large bearing on cross modulation distortion as will be discussed presently.

(i) Modulation envelope distortion

Because of the presence of the higher order terms in the power series representation, the magnitude of the r-f signal which can be handled by a particular type of r-f valve is limited if the distortion of the modulation envelope is not to be serious. Methods of measuring and calculating the signal handling capabilities of r-f amplifiers have been discussed in the literature and the reader is referred to K. R. Sturley's " Radio Receiver Design " Part 1 Chapter 4 Sect. 7 for a description of these methods, and to the other references listed at the end of this chapter.

(ii) Cross modulation distortion

Cross modulation distortion is an effect well known to receiver designers. It occurs when two signals are applied to any non-linear element such as a radio valve, under certain operating conditions, having appreciable third order curvature of the grid voltage-plate current characteristic. The effect is most noticeable when a strong local signal is present and the receiver is tuned to a weak signal. Insufficient r-f selectivity is often a major contributing factor in this regard. It is quite useless to have good selectivity in, say, the i-f amplifier once cross modulation has occurred in the r-f or converter stage, as no amount of subsequent selectivity can remove the undesired signal which is now superimposed on the same frequency band as the desired signal, and the modulation on both signals will be heard in the received output.

The r-f amplifier valve must also be designed so that the higher order modulation terms are kept as small as possible. A variable-mu valve will materially assist in handling a wide range of signal voltages without the grid voltage-plate current curvature being such as to allow serious cross modulation to occur.

A careful choice of the bias voltage for the r-f valve, and making the cathode by-pass capacitor of sufficiently small value to be effective at radio frequencies only, will often assist in reducing cross modulation.

When the local interfering station is a very powerful one it often becomes necessary to use special rejector circuits as well as high r-f selectivity.

Effects such as external cross modulation are not discussed here but reference can be made to F. E. Terman's " Radio Engineers' Handbook " page 647 for some details.

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